THE CROSLEY MODEL WLW SUPER POWER RADIO RECEIVER

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Bachelor of Science

In

Electrical Engineering

Oklahoma Agricultural and Mechanical College

1927

Submitted to the Department of Electrical Engineering

Oklahoma Agricultural and Mechanical College

In Partial Fulfillment of the Requirements

For the Professional Degree in

Electrical Engineering

1939

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PREFACE

Ordinarily, it would seem that the discussion of a radio receiver would occupy but small space. The most simple receiver of the superheterodyne type, however, requires an entire volume for a clear description if its different circuits are presented in a general as well as specific form. In describing a receiver as intricate and as complex as the one under consideration in this treatise, not one but several volumes would be needed for a comprehensive survey of all the circuits involved. Cognizant of this fact, but realizing too much space must not be used, the writer has presented a somewhat condensed treatise. Consequently, only the highlights have been touched upon.

The main body of the thesis has been divided into eight parts, with each part sub-divided into different topics. In general, the scheme is one in which the writer has endeavored first to present a general technical discussion followed by the specific example as to how it was incorporated in this particular assembly. Not all of the Mathematical analyses are original with the writer, but are, for the most part, revised by him and his colleagues of The Crosley Radio Corporation. All equations are workable to a fair degree of accuracy, and are handy tools for the radio engineer.

It is hoped that the sequence of presentation is found both logical and pleasing.

South Newport, Kentucky

March 1, 1939 A.P.R.

TABLE OF CONTENTS

PART I

HISTORY OF ORIGIN

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There is always some reason either logical or illogical, for any undertaking. In the engineering profession, if the engineer has any say, the reason is logical usually. The logic of Mr. Powell Crosley, Jr., President of The Crosley Radio Corporation, was not at once apparent when he gave orders for constructing the Model WLW Super-Power Radio Receiver in the early spring of 1936. As a matter of fact he was discouraged by several department heads and urged to forget the idea.

It is, however, characteristic of Mr. Crosley not to become discouraged easily. And furthermore he is a good salesman; enough so to win his point in an amiable manner. He referred to the Stratosphere Model of the Zenith Radio Corporation as an example of quality in radio receiver construction. But the Stratosphere Model made use of only twenty-five tubes and three loudspeakers! And Mr. Crosley, although he is a personal friend of Mr. MacDonald, President of the Zenith Radio Corporation, decided not only to equal attainments to date, but to surpass them by giving the world the largest and most powerful radio receiver yet known. Quality and richness of tone was to have been the best obtainable. It is needless to point out that his ownership of WLW, the world's most powerful broadcasting station, had no small bearing on the issue. He owned this station, why not go a step further and produce the world's

greatest radio receiver? He could and would!

With these receiver requirements as an objective, many engineering conferences were held. Invitations to attend these conferences were extended to the advertising, sales, cost, and purchasing departments. For an intelligent decision on accoustics and loudspeaker selection, Dr. Hugh S. Knowles, Chief Engineer, of the Jensen Radio Manufacturing Company, was called in as a consultant.

At last it was decided that such a receiver should incorporate no less than thirty tubes, six loudspeakers, four chassis, and a suitable cabinet.

From the engineering personnel, the writer was selected to undertake this (then fantastic) job. Although more intricate than receivers yet built, the work was a pleasure from beginning to end. There follows on the succeeding pages a description of the work.

PART II

GENERAL DESCRIPTION OF RECEIVER

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Any modern radio receiver can, for the sake of analysis, be broken up into four sections. These sections are classified as follows: variable radio frequency amplifier, fixed radio frequency or intermediate frequency amplifier, audio frequency amplifier, and power supply. This classification holds true regardless of the number of tubes used. In the receiver under description here, this general plan has been elaborated upon to the extent that it is unusual and more or less interesting.

Before further description is attempted, it must be understood that as good or better results could have been obtained by the use of other schemes than the one decided upon in this particular case. It also must be understood that the plan here adopted was necessarily in accordance with the wishes of Mr. Crosley. From the cost angle (engineers cannot ignore costs) it was perhaps the best plan, but from the angle of sheer engineering skill, it was not a desirable plan.

The WLW Model in its final form is, as has been pointed out, divided into four chassis, whose designations and functions are: (1) The L-1 Chassis containing the variable radio frequency or preselecting amplifier, the intermediate frequency amplifier, the pre-audio amplifier, and its own power supply. Within these four sections,

there are several features which will be described in detail under the heading of The L-1 Chassis, (2) The L-2 Chassis is essentially the power amplifier of the receiver as a whole. It is divided mainly into three frequency channels, which reproduces the entire audible range. These are, namely: the bass channel, the mezzo channel, and the treble channel. In addition to these three channels there is a fourth channel designated as the public address pre-amplifier channel. The reason for this latter channel being separate from all others, while not at once apparent, will be made clear in the detailed discussion of The L-2 Chassis.

(3) The L-3 Chassis. Little can be written of this and the (4) L-4 Chassis without a detailed explanation of each. Such a discussion will be given on pages to follow.

A schematic of the receiver in general is given in Fig. 1. Symbols representing the chassis and speakers are connected with lines, each line representing a circuit. Otherwise, the drawing is self-explanatory.

PART III

THE L-1 CHASSIS

TRIPLE TUNED INTERMEDIATE FREQUENCY TRANSFORMERS

The double-tuned intermediate frequency transformer of a superheterodyne receiver has two decided advantages over the ordinary tuned radio frequency transformer. They are, (1) excellent impedance matching networks making for high gain and (2) extreme sharpness of resonance. They may also be thought as a first class band pass filter network. For narrow bands to be passed, their performance is excellent and the lower the frequency at which they resonate, the better is both their gain and their selectivity. The very fact they have great selectivity or sharpness of resonance, gives rise to a major disadvantage when they are to be used in the reception of high quality radio programs where the band to be passed is from 12 kilocycles to 15 kilocycles. This statement is true even when an intermediate frequency of 450 kilocycles is employed. The practice of over coupling is frowned upon because double peaking of the resonance curve obtains and it cannot be controlled to any great degree of accuracy in production.

With these and many other problems in view, an attempt to obtain good band pass characteristics was made by introducing a third or tertiary circuit. It met with success, and the whole assembly became known as the triple tuned transformer. In the design of a triple tuned transformer, several factors must be considered. The two most important of these factors are Q (reciprocal of power factor) of the

coils or inductances and coupling. It must be admitted that practical design is largely a matter of cut and try, for a rigorous mathematical analysis of such a circuit is quite involved and to the writer's knowledge has never been published. In this paper, however, the writer will attempt an analysis by analogy. To do this it is first necessary to render a solution of the simple double tuned transformer.

As was stated above, a transformer may be considered as an impedance matching network, a general form of which is shown in Fig. 3. In this figure, Z_1 may or may not equal Z_2. The more general or unequal condition will be dealt with here. For this condition Z_{11} may be considered a generator impedance connected between terminals 1 and 2 and a load impedance Z_{12} connected across the output or terminals 3 and 4. The impedances looking in both directions at the input terminals 1-2 will be equal as is also the case in both directions at output terminals 3-4. These impedances are called the "image impedance" of the network and their values may be calculated in terms of Z_1, Z_2, and Z_3. By definition Z_{11} is the input impedance at terminals 1-2 when Z_{12} is connected across output terminals 3-4 and is

$$Z_{11} = Z_1 + \frac{(Z_2 + Z_{12})\, Z_3}{Z_2 + Z_3 + Z_{12}} \qquad (1)$$

Similarly, the impedance looking back from 3-4 with Z_{11} the generator impedance across terminals 1-2 is

$$Z_{12} = Z_2 + \frac{(Z_1 + Z_{11})\, Z_3}{Z_1 + Z_3 + Z_{11}} \tag{2}$$

clearing of fractions, equation (1) becomes

$$Z_{11}(Z_2 + Z_3) + Z_{11}Z_{12} = Z_1Z_2 + Z_2Z_3 + Z_1Z_3 + Z_{12}(Z_1 + Z_3) \tag{3}$$

and similarly equation (2) becomes

$$Z_{12}(Z_1 + Z_3) + Z_{11}Z_{12} = Z_1Z_2 + Z_2Z_3 + Z_1Z_3 + Z_{11}(Z_2 + Z_3) \tag{4}$$

Subtracting equation (4) from equation (3),

$$\frac{Z_{11}}{Z_{12}} = \frac{Z_1 + Z_3}{Z_2 + Z_3} \tag{5}$$

Adding equations (3) and (4)

$$Z_{11}Z_{12} = Z_1Z_2 + Z_2Z_3 + Z_1Z_3 \tag{6}$$

Multiply equation (5) by equation (6) and extracting the square root

$$Z_{11} = \sqrt{\frac{Z_1 + Z_3}{Z_2 + Z_3}(Z_1Z_2 + Z_2Z_3 + Z_1Z_3)} \tag{7}$$

Dividing equation (6) by equation (5) and extracting the square root

$$Z_{12} = \sqrt{\frac{Z_2 + Z_3}{Z_1 + Z_3}(Z_1Z_2 + Z_2Z_3 + Z_1Z_3)} \tag{8}$$

These image impedances Z_{11} and Z_{12} may be considered as pure resistances while Z_1, Z_2, and Z_3 may be considered as pure reactances of a T network. For a complete matching,

$$Z_1 = jX_1$$

$$Z_2 = jX_2$$

$$Z_3 = jX_3$$

$$Z_{11} = R_1$$

$$Z_{12} = R_2$$

are the elements of a complete network. The reactances may, of course, have either positive or negative values. Now substituting these values in equation (7)

$$R_1^2 = -\frac{X_1 + X_3}{X_2 + X_3}(X_1 X_2 + X_2 X_3 + X_1 X_3) \tag{9}$$

Similarly,

$$R_2^2 = -\frac{X_2 + X_3}{X_1 + X_3}(X_1 X_2 + X_2 X_3 + X_1 X_3) \tag{10}$$

If R_1 and R_2 are pure resistances, as stated above, the right side of equations (9) and (10) must be positive numbers and therefore one of the reactance arms must be opposite in sign to the other two arms.

By multiplying equations (9) and (10) and extracting the square root there is obtained,

$$R_1 R_2 = -(X_1 X_2 + X_2 X_3 + X_1 X_3) \tag{11}$$

Dividing equation (9) by equation (10)

$$\frac{R_1}{R_2} = \frac{X_1 + X_3}{X_2 + X_3} \tag{12}$$

Now, in case of transformers, X_1 and X_3 may be considered as one term and $X_2 + X_3$ as another. The reactance

X_3 is called the "Mutual reactance" because it is common to both the input and output circuits. Then let

$$X_1 + X_3 = X_p$$

$$X_2 + X_3 = X_s$$

$$X_3 = X_m$$

Substituting these new values in equation (11) it becomes,

$$R_1 R_2 = X_m^2 - X_p X_s \tag{13}$$

and equation (12) becomes

$$\frac{R_1}{R_2} = \frac{X_p}{X_s} \tag{14}$$

In design, one of the three arms may be selected arbitrarily and the other arms determined from equations (13) and (14). By combining these two equations, a set of equations may be obtained which gives each arm in terms of the others. Solving equation (14) for X_p and substituting its value in equation (13)

$$R_1 R_2 = X_m^2 - \frac{R_2}{R_1} X_p^2$$

from which

$$X_p = \pm\sqrt{\frac{R_1}{R_2}(X_m^2 - R_1 R_2)} \tag{15}$$

likewise,

$$X_s = \pm\sqrt{\frac{R_2}{R_1}(X_m^2 - R_1 R_2)} \tag{16}$$

Furthermore,

$$X_m^2 = X_p X_s + R_1 R_2 \tag{17}$$

$$X_m = \pm\sqrt{\frac{R_2}{R_1} X_p^2 + R_1 R_2} \tag{18}$$

$$X_m = \pm\sqrt{\frac{R_1}{R_2} X_s^2 + R_1 R_2} \tag{19}$$

By definition, $X_m = \omega M$ where ω is the angular velocity to $2\pi f$ and M is the mutual inductance of two coupled circuits. The coefficient of coupling is $k = \frac{M}{\sqrt{L_p L_s}}$ where L_p and L_s are the inductances corresponding to the reactances X_p and X_s.

In radio frequency transformers, k can never equal unity, and therefore it is impossible for $X_m \gg \sqrt{R_1 R_2}$ as is necessary in iron core transformers. It is, however, necessary to satisfy equation (15) and (16) and this condition is met when X_m^2 is equal to or greater than $R_1 R_2$. When $X_m^2 > R_1 R_2$, the condition is defined as sufficient coupling whereas when $X_m^2 = R_1 R_2$ the condition is called critical coupling. Since R_1 may be considered the resistance of X_p and R_2 the resistance of X_s, this condition of critical coupling is the key to real design of coupled circuits of which the double tuned intermediate frequency transformer is a splendid example.

Fig. 4 is a schematic of the double tuned transformer as it is used and Fig. 5 are curves showing the degrees of coupling. The curve A is obtained for the case of in-

sufficient coupling or where $X_m^2 < R_1R_2$, curve B shows the condition for critical coupling or where $X_m^2 = R_1R_2$ and curve C the condition of sufficient coupling where $X_m^2 > R_1 R_2$.

The five equations (15 to (19) inclusive are the fundamental equations used in the design of radio frequency transformers. An inspection of them, however, shows that they are not rigorous solutions of the problem. Another factor, important as it is, has been omitted purposely for the sake of simplicity and since it cannot be overcome physically, the solutions are accurate enough for commercial applications. This factor referred to is that of capacity coupling between the inductances themselves and between their leads. The effect of capacity is to lend asymmetry to the selectivity curve. It might be mentioned parenthetically that this condition has been improved upon of late by careful physical construction of the transformer. In spite of efforts made in this direction, it is doubtful that an absolutely symmetrical curve can ever be obtained because as the phase angle changes, at the point of inflection (peak), from lagging to leading the slope becomes different due to the very laws of inductance and capacitance. However, this condition is not noticeable at amplitudes greater than 10^{-3} times full amplitudes. Thus, it is not too serious.

Having obtained equations (15) to (19) inclusive, they may be applied to the design of a triple tuned transformer. To illustrate how this is done, reference is made to Figs. 6 and 7. In Fig. 6 is shown the circuit application of the transformer where T_1 is the input tube, T_2 the output tube, P the primary inductance, T tertiary inductance, S the secondary inductance, M_1 the mutual inductance between P and T, M_2 the mutual inductance between T and S, and M_3 the mutual inductance between P and S. If M_1 and M_2 meet the condition of critical coupling ($X_m^2 = R_1R_2$), M_3 must necessarily be very small or $X_m^2 \ll R_1R_2$ to obtain a curve like that of B, Fig. 7. Should M_1 and M_2 become greater, M_3 increases in a far greater ratio and a curve like that of C will obtain. It must be admitted that intelligent trail and error methods produces the final solution more rapidly than otherwise after the design has been based on the above equations.

The adjustment or alignment of a triple transformer gives proof physically of the equations used for its design. Assumption is made first that critical coupling for M_1 and M_2 has been established. The procedure then is to purposely detune the tertiary circuit and resonate the primary and secondary circuits to the desired frequency. This will result in curve A of Fig. 7, which is the condition for insufficient coupling where M_3 is very small.

The tertiary circuit is then brought into resonance and curve B results. These curves may be obtained either by point by point measurement or by oscillographic methods.

For the case of over coupling curve C will result as stated above. An examination of the peak or top of this curve brings out another important fact which is that a complete mathematical solution of the triple tuned transformer would be an equation of the sixth degree!

Analysis of the double tuned transformer has been attempted from other angles of approach, but none of them are so complete that the design engineer without experience can produce a commercially practical unit directly from such solutions. The triple tuned transformer is infinitely more difficult and as was stated previously there is no complete solution of such a circuit. Therefore, in designing three tuned circuits it is best to resort to knowledge gained from experience with the double tuned coupled circuits. Regardless of this handicap, the triple tuned transformer can be so constructed that a relatively flat topped wide peak is obtained with a resulting increase of the audio frequency range. This paves the way for high fidelity.

AUTOMATIC VOLUME CONTROL

Although the Automatic Volume Control has been used for several years and was not a new development for the receiver under discussion here, it is felt that a brief description of its function should be given. The reason for such an explanation will become apparent when other features involving the automatic control principle are

discussed.

The term automatic volume control is somewhat of a misnomer in its true function. A more lucid term to be applied would be automatic gain control. In reality its main action is to compensate the gain of the radio amplifier stages in accordance with field strength variation of a received signal.

Usually the radio frequency amplifier tubes are of the so-called variable mu type. They are designated as such because the peculiar design of their control grid permits a wide variation of transconductance versus control grid voltage. Transconductance, (S_m) formerly referred to as mutual conductance (g_m) is defined as the ratio of plate current change to grid voltage change. More strictly speaking it is the partial derivative of plate current with respect to grid voltage. The equation being

$$g_m = S_m = \frac{\partial i_p}{\partial e_g}$$

As is indicated, the function is not linear, although it is practically so over the narrow range of $e_g = 0$ to $e_g = -10$ volts. Complete plate current cut-off occurs for these tubes, when $e_g = -60$ volts approximately. With $e_g = -60$, $i_p = 0$, which results in $S_m = 0$ and since S_m is a measure of stage gain, the amplification is correspondingly low in all stages. Such a case is represented when a powerful local station is tuned in.

For a technical description, reference is made to Fig. 2. In this drawing, 1, 2, and 3 represent three radio frequency amplifier tubes with all elements omitted except the essentials; viz, cathode, control grid, and plate. This is done for the sake of simplicity. The radio frequency transformers are designated as, T_A antenna transformer, T_{RF} interstage transformer, T_{IF} intermediate frequency transformer, and T_D intermediate frequency diode transformer. The resistor R_D is referred to as the diode load resistor. It is across this resistor that the audio frequency voltage and the automatic volume control voltage is developed as will be shown later. The resistors R_1,R_2, and R_3 are termed isolating or filter resistors. Condensers designated as C are radio frequency by-pass condensers of value such that their reactances are negligible at the frequencies involved. The condenser C_D is known as the diode rectifing condenser. More prpperly it should be termed the radio frequency by-pass condenser.

A mathematical analysis of the diode as a linear detector and rectifier will not be given here, because other works have given an analysis in quite some detail. Equations will be given, however, to show what relationships exist in such a unilateral device.

With the system shown in Fig. 2, tuned to a resonance with the incoming signal, the instantaneous voltage applied to the diode is given by the equation

$$e_p = E' \cos \theta - E_a \tag{20}$$

where e_p is the applied instantaneous voltage, E' maximum radio frequency voltage, $\theta = \omega t$, and E_a the direct current component of voltage. At some angle θ_1, e_p becomes equal to zero and

$$\cos \theta_1 = \frac{E_a}{E'} \tag{21}$$

The direct current component I_o can be calculated from the relation

$$I_o = \frac{E_a}{R_D} = \frac{E' \cos \theta_1}{R_D} \tag{22}$$

It can be shown also that

$$\frac{R_P}{R_D} = \frac{1}{\pi} (\tan \theta_1 - \theta_1) , \tag{23}$$

Where R_p is the plate resistance of the diode.

Such is the relationship between the angle θ_1 (usually called the operating angle) and the ratio of the diode plate resistance and its external load resistance. This function is a transcendental equation for which there is no direct rigorous solution. Graphical solutions are practical, however, and solving for the reciprocal of equation (23), the value of E_a can be obtained. The voltage represented by E_a is, of course, the automatic volume control voltage applied to the control grid tubes 1,2, and 3 through resistors R_1, R_2, and R_3.

Furthermore, it can be shown that

$$E^{M}_{aud} = m\,E'\left(\frac{E_a}{E'}\right) \qquad (24)$$

Where E^{M}_{aud} is maximum or peak audio frequency voltage developed, and m the modulation factor of the radio frequency voltage. It is thus seen that one diode may be used to obtain both the audio frequency voltage and the Automatic volume control voltage.

Reference is again made to Plate I to show how the automatic volume control was obtained in the Model WLW receivers. Attention is first directed to the intermediate frequency transformer 6 whose tertiary winding (shown here as the center winding) energizes the grid of tube 73C. The plate circuit of this tube has for its impedance the primary of intermediate frequency transformer 8 whose secondary is center tapped. The ends of this winding are connected to diode plates P_1 and P_2 of tube 72, while the center tap is connected through resistors 54A and 49 to ground. It is across these two resistors that the automatic volume control voltage is developed. By following the other lead from the center tap, it can be seen that full automatic volume control voltage is applied to the grids of tubes 73A and 74 through the filter resistors 73 and 71. Only a portion of the full automatic volume control voltage is applied, however, to the grid of tube 73B. At the junction of resistors 54A and 49, resistor 52B connects the "low" side of the true secondary of

intermediate transformer 6 which in turn leads to the grid of tube 73B. This control voltage is dependent upon the total voltage developed across the resistors 54A and 49 and the ratio to each other. The relation that exists is

$$E_1 = E_o \frac{R_2}{R_1 + R_2} \tag{25}$$

where E_0 is the total voltage developed, R_1 is resistor 54A, R_2 is resistor 49, and E_1 the voltage applied to tube 73B. Since 54A equals 1 megohm, and 49 equals 150,000 ohms equation (25) reduces to

$$E_1 = E_0 \frac{.15}{1 + .15}$$

$$E_1 = .13\ E_0$$

Thus, it is seen that slightly more than one-eighth of the total voltage developed is applied to the control grid of tube 73B. Such practice is usual, for with present day tubes full voltage does not need to be applied to more than two tubes. The best practice is that of applying full voltage to the first tube and tapering the voltage for subsequent stages. Unfortunately, however, there are commercial limitations to be observed.

The intermediate frequency transformer 8 has another function which will be discussed in another part of this paper.

TUNING INDICATOR

The tuning indicator functions as a means of informing the operator of a radio receiver whether he is properly tuned to the desired transmitter. It usually operates from the d-c voltage developed across the diode load and its amplitude is directly proportional to the field intensity of the carrier frequency. Many styles and types have appeared. The earlier types were of the d-c meter origin with very cheap parts and of still cheaper construction. These facts cannot be helped, because in a commercial radio, economy is not only a by-word, it is the watchword! Indicators of this type were not entirely satisfactory, however, because the movement was highly damped. To the layman, it appeared sluggish while tuning. Then the moving vane type was developed and held sway for a number of years. In its wide variety of styles, it was more satisfactory than the first type.

The most notable advancement in the art of tuning indicator came when RCA announced the "Magic Eye". This type of indicator operates on the principle of the cathode ray tube and is more technically known as the cathode tube electron-ray indicator, which is especially designed to give visual indication of voltage changes. The elements of such a tube are (1) heater which heats the cathode to a temperature sufficient for emission, (2) cathode-electron emitter, (3) target, flourescent coated upon which the electrons impinge, (4) plate, providing potential differences between

it and cathode essential to give electron velocity necessary for them to strike the target, (5) control grid, to which is applied changing voltages. In actual construction, the elements are so designed that as changes in voltages are applied to the control grid, the illuminated area of the target varies. Furthermore, the design is such that the illuminated area becomes greatest with the greatest negative voltage applied to the control grid. It is obvious from the discussion of automatic volume control that if the voltage is applied to the control grid of the indicator tubes, a visual indication of whether the transmitter is tuned in can be obtained. Two designs of target are available; viz, one which gives angular indication and the other operating as the iris diaphram of a camera shutter. This electron type is the most desirable on the open market, since, of course, the electron stream has no appreciable inertia and no lagging action.

Something different; something original, however, was demanded for the Super-Power Receiver. After much study, trial, and error, a workable scheme evolved. Since the TRADE-MARK of the Crosley Radio Corporation is the name Crosley with the symbol of lightning drawn through it, the natural impulse was to provide a realistic lightning flash through the name as a station was tuned in. The idea was approved and preparation made to incorporate it.

The means for illuminating was accomplished by using a special kind of neon tube as 92 in Plate I. The tube

socket has flexible leads coded black, red, and green for convenience. When the tube is inserted into the sockets, these leads makes connections to the tube as follows: black to low d-c potential or chassis, green to high d-c potential striking voltage, and red to plate of its indicating voltage amplifier tube designated as 78A.

In operation the striking voltage applied across terminals black and green breaks down the gas insulation at the base of the tube causing ionization of the neon which is manifested as a small glow. With no signal introduced this small glow is all that occurs for normally very little voltage is applied to the red lead. The reason for this phenomenon is that tube 78A, which is in effect a d-c amplifier, has no bias voltage applied and would draw approximately 15 milliamperes but for the series dropping resistor designated as 68 which has a value of 30,000 ohms capable of dissipating four watts. Actually, however, about eight milliamperes of plate current flows through the 30,000 ohms resulting, by Ohm's Law, in a voltage drop of 240 volts. Since that value is practically the potential available, there is little or no voltage applied to the red terminal of the neon tube.

Upon tuning in a station, a negative voltage is applied to the grid of tube 78A which is connected to a diode load 48B through the two megohm resistor 56. When about minus eight and one-half volts has been applied to the grid of tube 78A, plate current cut-off occurs and only a very small

current flows through resistor 68. Now, since this current is, in magnitude, a few hundred microamperes, practically all of the 240 volts potential available is impressed across the red and black elements of the neon and filling the entire tube with an orange-red glow.

To gain the effect desired the neon tube is placed behind a dial mask in which, through the name of Crosley, a jagged slit symbolic of lightning has been cut. This lightning flash which occurs when a station is tuned in, has a very pleasing effect upon the novice. It must be pointed out, however, that the flow is steady as long as a station is tuned in and its field remains great enough to develop the proper diode voltage. When the field strength decreases, or "fades" sufficiently, the glow necessarily wanes. Admittedly, this type of indicator falls short of the accuracy which can be obtained by other methods. In a spectacular way, it leaves nothing to be desired.

AUTOMATIC FREQUENCY CONTROL

Automatic frequency control has in the past been referred to also as automatic tuning, but with the advent of motorized and push button tuning the term automatic tuning adds confusion. The two are separate and distinctly different during the present state of the art, and hence should not be used synonymously. Automatic tuning is now understood to mean closing a switch or rotating a dial to tune in the desired station. The term automatic frequency control is construed to mean the ability of the receiver to remain in exact resonance with the transmitter frequency despite shifts

in circuit constants due to thermal, humidity, and other conditions. With the two terms defined, there follows a discussion and explanation of the automatic frequency control circuit and its application to a superheterodyne receiver.

The very fact that a receiver contains an oscillator and one or more stages of tuned coupled circuits, makes it highly susceptible to small changes in circuit parameters. The major change is frequency drift, and the most important factor contributing to this condition is thermal changes in the chassis upon which the component parts are mounted. The power transformer is of course the largest single unit source of heat dissipation. The rectifier and power output tubes are the next largest sources. In the case of this particular chassis, there may be as much as 300 watts or 350 watts heat dissipation, a large amount of which is transfered by convection and conduction to the component parts. Frequency drifts as high as 25 kilocycles have been measured in the regular broadcast band on some rather poorly designed chassis. Drifts of 5 to 8 kiloeycles are quite commonplace. The necessity for some sort of frequency stabilization is, therefore, apparent. To make it a commercial success, it had to be automatic in its action, thus requiring no attention on the part of the owner of the receiver.

Automatic frequency control systems have been employed in telephone carrier circuits for a number of years, but

their application to radio receivers would have been prohibitive from the cost standpoint. One system was developed at The Crosley Radio Corporation. It was a laboratory success, but proved to be rather crude because it was an electro-mechanical device requiring close mechanical adjustment. Therefore, it was abolished. The electronic system about to be described, was developed by The Radio Corporation of America. Only electrical adjustments were required.

Since the oscillator circuit of a superheterodyne receiver is the determinant of the dial calibration, because it combines with the incoming frequency to establish the intermediate frequency to which the intermediate frequency transformers resonate, and furthermore since it is most susceptible of all circuits to thermal changes, it is the logical place to which automatic frequency control should be applied. Fortunately it is the most easily controlled circuit of them all in a receiver. There follows a description of such a system.

An Automatic frequency control system must be divided into two parts or modes of operation. These are namely, the discriminator circuit and the control circuits. In a superheterodyne circuit, the incoming frequency and the oscillator frequency are fed into a common mixer tube, The output of the mixer tube consists of a single radio frequency (referred to as the intermediate frequency) modulated, of course, at an audio rate. This resultant

frequency is the difference between the oscillator frequency and the incoming frequency. For example, if the incoming frequency is 1000 kilocycles and the oscillator frequency is 1450 kilocycles the resultant or intermediate frequency is 450 kilocycles and remains such for all values of incoming frequencies. This is true, because the receiver is so designed that the difference between the incoming and oscillator frequencies remains constant. Mathematically it is expressed as,

$$f_{osc} - f_{in} = \text{a constant.}$$

Now let it be assumed that in the example given above, the oscillator frequency were actually 1453 kilocycles instead of 1450 kilocycles as the design had determined it should be. The frequency difference in this case would be 453 kilocycles. A further assumption is that the intermediate frequency transformers are tuned to resonate at 450 kilocycles. This means t that in effect the incoming signal frequency is mistuned by 3 kilocycles. Such a value of mistuning is quite likely in most cases and can be caused by either oscillator drift, or by inaccurate setting by the operator. It is an established fact that very few people, indeed, tune their receivers to exact resonance with the incoming signal. All forms of distortion thus enter and ruin any attempt at high fidelity reception. This, of course, applies to the receiver, not equipped with automatic frequency control. With automatic frequency control, however, the oscillator frequency would be automatically adjusted so that the receiver would be tuned within a very small percent deviation of exact resonance.

The discriminator circuit of the AFC (automatic frequency control) is based directly upon the principle of mistuning. To recapitulate, it must be remembered that the automatic volume control voltage is obtained by the use of a diode rectifier. In obtaining the AFC voltage, use is made of a differential diode rectifier.

The action depends upon the fact that a 90 degree phase difference exists between the primary and secondary potentials of a double tuned, loosely coupled transformer when the resonant frequency is applied and that this phase angle varies as the applied frequency varies. Thus if the primary and secondary voltages are added vectorially, the obsolute magnitude of the resultant vector will be greater on one side of resonance than on the other.

The vector sum of the primary and secondary voltages may be physically realized by connecting the two parallel tuned, coupled circuits in tandem applying the input potentials to one circuit and taking the output across both circuits in series. In this manner, an action similar to that of a side circuit is produced even though the primary and secondary are both tuned to the center frequency. See Fig. 8. Notice that the difference between A and B is the sign of the coupling between primary and secondary of the i-f transformer. The potentials at either end of a secondary winding with respect to a center tap, rather than one end of the secondary is connected to the primary, two potentials may be realized, one maximizing above and one maximizing below the center frequency. See Fig. 9.

If a transformer is connected in this manner and the resonant frequency is applied to the primary the two resulting output potentials will be equal in magnitude. If these are then applied to two separate, like detectors, and the resulting d-c voltages (or currents) are added in opposition, the sum will be equal to zero. If, however, the applied frequency departs from resonance, the sum of their outputs will be some real value whose polarity will depend upon the sign of the frequency departure.

The rate of change on scalar magnitude of a given resultant of two vectors at 90 degrees, with small changes in the angle between those vectors, is greatest when the scalar of one vector is equal to the scalar value of the resultant divided by $\sqrt{3}$, or when the ratio of vector lengths is equal to $\sqrt{2}$. If a double tuned transformer has a secondary of twice the inductance of the primary, the Q of the primary being equal to that of the secondary (when in circuit) and the coupling between circuits being critical, the primary voltage will be related to one-half the secondary voltage on resonance in such manner as to fulfill the above conditions.

This does not mean that a larger secondary with the same primary, or a different value of coupling, would not give a greater number of volts per cycle change in the primary plus $\frac{1}{2}$ secondary sum, but in such event the resultant itself would be greater. Circuit or other requirements might necessitate an exceedingly low tuned primary impedance in which case a much higher ratio would be in order.

A measure of the sensitivity of this device may be the developed d-c volts (or amperes) per cycle of frequency deviation, per volt applied to the grid of the tube whose plate circuit contains the primary of the transformer. Regardless of the type of detectors employed this quantity will be a function of the rate of change, with frequency of the difference between magnitudes of the input potentials to the two detectors. If these magnitudes are plotted against frequency difference (both positive and negative) the curves will intersect on the zero abscissa ordinate with slopes equal but opposite in sign. See Fig. 9. The slope of the curve representing their difference is, therefore, equal to twice the slope (at the center frequency) of the curve of input potentials to one of the detectors. This establishes the significance of a factor which will be termed S, which equals two times the first derivative, with respect to frequency, at resonance, of an expression for absolute magnitude of input potential to one of the detectors. It must be borne in mind that the value of the ordinate at the point of intersection of the two curves become significant only when detectors other than those with linear characteristics are used.

To simplify the derivation given below, the apparent Q values of both primary and secondary (when in circuit) have been assumed to be equal.

SYMBOLS EMPLOYED:

S = Slope at resonance, of the expression representing the

difference between magnitudes of the potentials applied to the two detectors.

f' = A frequency removed from f by a discreet increment.

r = Apparent primary series resistance. This includes the effect of the plate impedance of the tube, the natural primary series resistance, and any other resistive load other than the secondary.

r_2 = Apparent secondary series resistance.

A = The ratio of total secondary inductance to primary inductance.

Since $Q_1 = Q_2$ (by assumption) $r_2 = Ar$.

L = Primary inductance. Thus $L_2 = AL$.

$Q = \frac{2\pi fL}{r}$

X = Sum of the internal series reactances of the primary.

$X_2 = Ax$ = Sum of the internal series reactances of the secondary.

n = The ratio of reactance to resistance (internal) of either primary or secondary at any frequency. At f' this equals $\frac{2(f'-f)Q}{f}$.

K = The ratio between actual and critical couplings between primary and secondary $2\pi fM = K\sqrt{r}\sqrt{r_2} = K\sqrt{A}\ r$.

G_m = Mutual conductance of the amplifier tube preceding the transformer. $j = \sqrt{-1}$

With 1 volt applied to the grid of the amplifier tube the vector primary voltage takes the form

$$*E_p = 2\pi fLQG_m \frac{(1 + jn)}{(1 + jn)^2 + K^2} \tag{26}$$

The primary L current may then be written:

$$*I_p = \frac{E_p}{j2\pi fL}$$

And the induced voltage in the secondary equals $\frac{E_p K \sqrt{A}\, r}{2\pi fL}$

From which the secondary current becomes:

$$I_s = \pm \frac{E_p K}{2\pi fL(1 + jn)\sqrt{A}}$$

And the secondary voltage may be written:

$$*E_s = \pm \frac{jE_p K \sqrt{A}}{(1 + jn)}$$

Replacing E_p with its equivalent from (26):

$$E_s = \pm 2\pi fLQG_m \frac{jK\sqrt{A}}{(1 + jn)^2 + K^2} \tag{27}$$

Which is the expression for the vector voltage across the entire secondary with one volt applied to the grid of the amplifier tube.

Adding the primary voltage to one-half the secondary voltage vectorially, gives the vector expression for the resultant voltage to either detector.

Note * To a very close approximation.

$$E_{det} = 2\pi fLQG_m \frac{(1 + jn) \pm jK\sqrt{A/2}}{(1 + jn)^2 + K^2} \qquad (28)$$

or:

$$E_1 = 2\pi fLQG_m \frac{(1+jn) + jK\sqrt{A/2}}{(1+jn)^2 + K^2}$$

and:

$$E_1 = 2\pi fLQG_m \frac{(1+jn) - jK\sqrt{A/2}}{(1+jn)^2 + K^2}$$

The scalar magnitude of E_1

$$Y = 2\pi fLQG_m \frac{\sqrt{1+n^2 + \sqrt{A}\,Kn + AK^2/4}}{\sqrt{1+K^4 + n^4 + 2K^2 + 2n^2 - K^2n^2}} \qquad (29)$$

As stated above the sensitivity of the device to changes in frequency will be a function of S which is two times the rate of change of the above expression, with respect to frequency. Or $S = 2(dy)/df$. From the definitions above it can be seen that, at resonance, $dn/df = 2Q/f$. Thus $(4QdY)/(fdn) = S$. Differentiating Y with respect to n and then setting n equal to zero to get the slope at resonance gives:

$$\frac{dY}{dn} = 2\pi fLQG_m \frac{\sqrt{A}\,K}{(1+K^2)\,(1+AK^2/4)}$$

from which:

$$S = 8\pi LQ^2G_m \frac{\sqrt{A}\,K}{(1+K^2)\,(1+AK^2/4)} \qquad (30)$$

From this expression it can be seen that S is independent of frequency and is proportional to L, Q^2, and to G_m and also that it is a function of A (the secondary to primary inductance ratio) and of K, $\left(\frac{\%\ \text{critical coupling}}{100}\right)$.

If the right hand side of the expression is maximized by differentiating with respect to K setting the differential equal to zero and solving for K in terms of A, there results:

$$K = \frac{\sqrt{1 + 2A} - 1}{A} \tag{31}$$

from which it can be seen that the optimum value of coupling will be less than critical for any ratio of secondary to primary inductance. K is plotted against A in Fig. 10.

If the expression for S is maximized with respect to A, we arrive at a ratio equal to infinity with zero coupling. This merely confirms the fact that the sensitivity can be increased by increasing the secondary inductance. It must, of course, be borne in mind that if conductive input detectors are used, their effect on the apparent Q of the tuned circuits will be greater, the greater the inductance.

As an example of the use of (30), possible values for the parameters may be taken as follows: $L = .5 \times 10^3$, $Q = 100$, $G_m = 1500 \times 10^{-6}$; $A = 2$. Then from (31): $K = .785$. Substituting these values in (30) and solving for S we have: $S = .113$ rms volts difference, per cycle, in the potentials applied to the two detectors when one volt rms is applied to the grid of the preceding amplifier tube. Thus, if the frequency departs from resonance by 10 cycles an unbalance of 1.13 volts will exist at the detector input points. A sensitivity of this order is not, in gen-

eral, necessary or desirable. However, the above example illustrates the order of sensitivities which may be obtained should the need arise. The calculated value for S has been varified experimentally.

So far the slope has been calculated at resonance only. If the scalar magnitudes of E_1 and E_2 are calculated for positive and negative values of n, subtracted and plotted against n, it is apparent that the slope becomes equal to zero at two points. These correspond to two frequencies, one above and one below the resonant frequency, and at these points the difference between the applied potentials to the two detectors is maximum. With various circuit constants and with the coupling adjusted to give a maximum slope at resonance, these maxima will normally appear at positive and negative values of n ranging between .5 and .95. The frequencies corresponding to these values are sufficiently well separated to give adequate operating range for most applications. This is particularly true if the range of frequencies applied to the device is limited by the selectivity of preceding circuits. It must also be remembered that the differential d-c output voltage will bear the same sign after passing a maxmum point, and if it is used for frequency control it may still have sufficient magnitude to swing the controlled frequency into the so-called operating range.

However, if it becomes necessary to increase the frequency separation of the two maxima, it may be done either by increasing the value of the coupling above the optimum

as determined by (31), or by decreasing the Q of the circuits. Either method will decrease S at the center frequency, although an increase in coupling will cause the least change in sensitivity for a given increase in separation.

So far no mention has been made of methods for combining the d-c output potentials (or currents) of the detectors to produce the differential effect. A simple voltage connection will be described in detail.

Referring to Fig. 11, circuits I and II, both tuned to the same frequency, are mutually coupled and connected together as described above. The reactance of the condenser C_3 between points C and D is small at the frequency of operation and merely serves to isolate the d-c plate potential of the primary. The diodes connected with their plates at points A and B are conventional except for the fact that for this circuit they must have separate cathodes. The two diodes in a type 6H6 tube fulfill this condition. The diode cathodes are connected together by means of the condenser C_4 and one of them is also connected to ground. The condenser C_4 must have low impedance at the operating frequency and in general it will be desirable that it be low at useful modulating frequencies. Two resistors R_1 and R_2 are also connected, in series, between cathodes. Their resistances are equal and will usually be between .5 and 1.0 megohm. The center point F between them is connected to the center tap C on the secondary. The use

of an r-f choke in this connection is optional but if it is used the condenser C_5 (shown dotted in the diagram) must also be included and the two together will serve to decrease the effect of the resistors on the Q value of the primary.

The action is as follows. If the resonant or "center" frequency is applied to the grid of the amplifier tube, equal amplified voltages will exist between the point A and ground and between the point B and ground. These are rectified by the diodes and direct currents will flow in the resistors R_1 and R_2 in opposite directions with respect to ground. Thus, the net d-c potential produced by the two IR drops between E and ground is equal to zero. If, however, the applied frequency departs from resonance the potentials across the diodes will be unequal in magnitude, unequal IR drops will be produced in the two resistors and d-c potential will exist between E and ground, the polarity of which will depend upon the sign of the frequency departure.

As has been indicated on the diagram, a-f and AVC (automatic volume control) voltages also may be derived from the rectified output of this circuit, as well as the differential d-c potentials.

If a carrier at the resonant frequency with normal "intensity" modulation, but without frequency modulation, is applied to the system, the a-f as well as the d-c voltages across R_1 and R_2 will be equal and opposed. Therefore, at resonance there will be no a-f potentials between E and

ground, and as far as audio components are concerned the system acts exactly as though point E were grounded with the outputs of the two diodes acting in parallel. Actually if C_4 is sufficiently large to have negligible reactance at the lowest modulating frequency, this is the case. Then the point F becomes a potent source of audio voltages to supply the a-f amplifier system and no other audio detector is necessary. If AVC voltages are also taken from the point F and it is necessary to maximize the ac-dc impedance ratio, it can be seen that the d-c impedance is equal to one-half the resistance of one of the resistors even though R_1 and R_2 are not in parallel as far as d-c is concerned. The use of a normally active control element in an automatically controlled i-f frequency system will not allow the carrier to depart sufficiently from resonance to hazard the above facts.

It can be seen that the d-c potential between ground and the point F will have the proper polarity to be used for AVC action, and that this potential will bear the same ratio to the developed audio voltages as is found in the conventional detector AVC system. The fact that it maximizes at one side of resonance is of no significance if automatic frequency control is used. When the AFC is cut out of the circuit (manually) it is assumed that the point E will be grounded. This will cause the d-c potential at F to maximize on resonance.

The only factor which determines the polarity of the AFC differential voltage developed at point E in Fig. 11 is

the sign of the coupling between the two coupled tuned circuits. An examination of the circuit shows that any unbalance in the capacities from either diode plate to ground takes on the nature of capacity coupling between the tuned circuits. Likewise an unbalance of capacities from either side of the secondary tuning condenser to ground has the same effect. Because of the mechanical construction of the average i-f transformer, this unbalance will usually be sufficiently large to account for the major portions of the coupling when it is connected in circuit as in Fig. 11. For this reason it is best to phase the inductive coupling to oppose this external capacity coupling (regardless of the d-c polarity requirements) and to correct the overall coupling by the addition of a small capacity from the proper diode plate to ground so as to either increase or decrease the existing capacity unbalance as the need may be. The d-c polarity of the AFC voltage may be reversed, if necessary, by removing the ground from the one diode cathode and grounding the opposite one.

It should be noted that the "transformer" gain (audio output + modulated i-f input) of the system in Fig. 11 will be greater than with conventional circuits since the primary and one-half secondary voltages are added (vectorially), and that the selectivity will approximate that of the primary alone.

An extension of the analysis reveals (as would be guessed) that modulation envelope distortion arises from the asymmetrical frequency discrimination. As expected, measurements shows this distortion is not appreciable at low

a-f modulating frequencies, but that when S is desirably great, I and II being coupled to a desirable degree, this distortion is appreciable when the carrier is deeply modulated (say 80 percent) at a frequency higher than (say 3500 cycles). Of course deep modulation does not normally occur at high audio frequencies, but as the frequency is increased a lesser percentage modulation gives rise to distortion which is not unquestionably negligible. Consequently, it cannot be safely recommended that the a-f output be taken across R_2 (Fig. 11) in a strictly high fidelity receiver.

As shown in Fig. 12(b), both AVC detection and a-f detection may be applied at the primary (circuit), and of course the same or separate diode (s) may be used.

Fig. 12(c) shows another modification of Fig. 11. The AVC point is tapped down on R_2. As best determined by experimental development, performance may in some cases be definitely improved by the tapped R_2 arrangement in Fig. 12(c). In Fig. 11, the AFC/AVC d-c voltage ratio does not exceed unity. In Fig. 12(c), unity may be exceeded when desirable.

In Fig. 12(a) and 12(b), of course C_o should be small (40 or 50 mmf), and the ordinarily high Q choke L_o must be large enough for $\omega L_o > 1/\omega C_o$: e.g., the frequency of series resonance of C_o and L_o should be of the order of I.F./10. Also L_o should be so chosen that it does not establish parallel resonance with the diode plate;

cathode capacitance(and circuit capacitance) at or near the I.F. To a first approximation, C_0 and C_R are the capacitances across R at a-f., from the standpoint of modulation envelope detection.

In order to make use of the d-c voltage differences developed by the discriminator circuit, they must be applied to some type of control circuit. Such a control must operate upon the tank circuit of the oscialltor to produce frequency changes. The control circuit, therefore, resolves itself into a variable reactance. This conversion of discriminator voltage differences is accomplished by means of a vacuum tube, which is termed the control tube. An analysis of the control tube in the role of a variable reactor follows.

Consider the circuit of Fig. 13 shown in simplified form in Fig. 14. Here E is tank circuit voltage.

i_1 is current in circuit R_1 C_1.

G_m is mutual conductance of control tube T_2.

i_p is a-c plate current of T_2.

e_g is a-c grid voltage of T_2.

Z_o is effective impedance of T_2.

A high impedance r-f pentode will ordinarily be used for the control tube so r_p can be neglected. The resistance of L_1 can also be neglected.

Then

$$i_1 = \frac{E}{R_1}$$

since in practice $R_1 = \frac{1}{j\omega C_1}$

$$e_g = \frac{i_1}{j\omega C_1} = \frac{E}{j\omega C_1 R_1}$$

$$i_p = e_g G_m = \frac{EG_m}{j\omega C_1 R_1}$$

$$Z_o = \frac{E}{i_p} = \frac{j\omega C_1 R_1}{G_m}$$

Since Z_o varies directly as frequency it has the nature of inductance. If L_o is called the virtual inductance due to the control tube,

$$L_o = \frac{C_1 R_1}{G_m} \tag{32}$$

If an inductance L_A were used in place of C_1,

$$Z_o = \frac{R_1}{j\omega L_A G_M}$$

In this case Z_o is effectively a capacity since it varies inversely as the frequency.

The use of capacity in the grid of the control tube has several advantages:

1. The Q of condensers is generally higher than that of inductances so that the control tube acts as a more nearly pure reactance.
2. The distributed capacity frequently resonates an inductance within the frequency band used, so that the control action disappears at that frequency.

3. A capacity appears as an inductance in parallel with the tank circuit inductance so that the frequency shift is a constant percentage of the resonant frequency throughout the tuning range.

In the circuit of Fig.13 the padding condenser, C_2 is placed at the high potential side of the circuit so that the control tube may be connected directly across L_1. If the control tube is placed in parallel with L_1 and C_2 in series, a certain amount of control is lost at the low frequency end of the band. The combination of L_1 and C_2 is resonant below the band, and at such frequency the control tube could have no effect. The circuit of Fig. 8 shows a blocking condenser C_4 connecting the low side of the tank inductance to ground, so that the plate voltage may be applied to the control tube through L_1.

Choice of the control tube is somewhat limited by two very important factors; (1) magnitude of discrimination voltage developed, and (2) mutual conductance of the tube itself. The discrimination voltage demands a tube with a quick cut-off characteristic, that is; one whose plate current is very low for a negative grid bias not to exceed seven or eight volts. The mutual conductance must be relatively high and also must change rapidly with grid voltage changes. Such conditions are met with admirably in the 6J7 tube. This tube is thus almost universally adapted to control functions.

Attention is again directed to Plate I for a description of the AFC system developed for this particular receiver. In this circuit, tube 73C is the discriminator frequency amplifier and is energized by the tertiary winding of the triple tuned transformer 6. This tube energizes the discriminator transformer 8 whose secondary terminals are connected to the differential diode rectifier tube 6H6 marked 72. Cathode K_1 of 72 is connected directly to ground. Cathode K_2 from which is developed the AFC voltage, connects through resistors 57 and 52D with the control grid of 6J7 tube marked 76. This tube (6J7) is the control AFC tube and its plate operates upon the oscillator tank circuits 9, 10, and 11, in parallel with a leg composed of condenser 23A and choice of resistors 44, 45, and 47 which are selected according to the frequency band in use. It will be noticed that this particular circuit differs slightly from that described in general on the fore-going pages. They function in a similar manner, however, except the latter is an improvement over the more simple type described in detail.

FIDELITY CONTROL

At the time this receiver was designed, fidelity control of several positions was in vogue. Positions of the control described herein are as follows:

A. Receiver off

B. Normal

C. High Fidelity

D. Mellow tone

E. Bass

F. Noise Reducing

The complete circuit including switching is shown in Plate I. These circuits are rather difficult to follow so the writer has redrawn their essentials in Fig. 15, with B to F inclusive in the same order as given in the above list. With the switch thrown to position B, the circuit in Fig. 15B obtains. Now, if an a-c voltage, constant in magnitude but varying from 30 cycles per second to 10,000 cycles per second, is impressed across the input circuit as shown in Fig. 15B, an output voltage of magnitude e_2 will result. (In general the magnitude of e_2 is plotted and shown as a curve in Fig. 16. The curves are designated as B, C, D, E, and F to correspond with the circuits in Fig. 15 which produce them. The value of e_2 is plotted as deviation in decibels from the 400 cycle response.)

The audio response as given by curve B, Fig. 15, is typical of that obtained with receivers of earlier design. Only the middle range of audio freqnencies are passed

through the circuit network and reproduced. The low and high frequencies are both rejected. It represents more or less the frequency response of the average telephone circuit. This position is used advantageously for the reception of speech.

The circuit network shown in Fig. 15C gives the response of curve C in Fig. 16. Such curve is called a sway-backed response. It will be noticed that the low frequency response is brought up to about + 17 decibels at 40 cycles and the 4000 cycle response is about + 6 decibels with respect to that at 400 cycles per second. This is designed to compensate for what is known as the Fletcher Effect of the human ear, which shows that the normal ear discriminates against both the high and low frequencies of the audio range. To the ear such compensation results in an apparent flat response from 40 cycles per second to about 7,000 cycles per second. It is known commercially as the high fidelity response. Fairly decent reproduction of musical programs can be had with such a response.

From the network of Fig. 15D, there is obtained the response curve D in Fig. 16 which is designated as Mellow Tone. The easiest way in which to describe its quality of reproduction is to say that it sounds as though it were issuing from within a large barrel. An inspection of curve D shows that the low and middle frequency response is comparable to that obtained through the high fidelity

network (curve C). The high frequency response is, however, greatly suppressed. The essentials for speech reception are present if one does not mind the accentuated bass.

A large number of consumers desires the "boom-boom" type of reception; especially so for dance programs. Hence, the network of Fig. 15 E was evolved and is called the Bass response. It will be noticed that C_5 and R_4 of this network contribute greatly to the cut-off of high frequencies.

The Noise Reducing response, curve F Fig. 16, is obtained through the network shown in Fig. 15B with the exception that it has better high and low frequency responses. This circuit is an aid to reception through static or other electrical interference and is used advantageously in downtown localities or during all but severe electrical storms.

In the circuits of Fig. 15, all parameters are constant except those designated by letters with subscripts. A change in the subscript denotes a change in value of the circuit element. This gives an easy method for determining just what elements are involved in changing from one circuit to another.

It might be added that the fad for multiposition fidelity control seems to have existed only during the year of 1936. At the present time, the fidelity control is of the continuously variable type, making possible an infinite

number of positions instead of only five.

AUTOMATIC VOLUME EXPANSION

During the rendition of a musical selection by orchestra or band, the sound intenisty (level, volume, etc.) may vary from that which is barely audible to that which assumes ear shattering proportions. Measurements with sound equipment show actual differences in level as much as 70 decibels. Such changes in volume are necessary to render the effect intended by the composer. This is all very well when the selection is not being broadcast. In broadcasting musical selection, however, several problems are encountered.

The first of these is presented by the transmitter itself. In order to assume reasonable area coverage, it is the practice to modulate at a fairly high average percent at all times, the greatest peak variation being perhaps from 10% to 100%. These two values represent a numerical ratio of ten to one in modulating factor. But this means a change of only 20 decibels! The other fifty decibels change (considering maximum crescendos) is lost. Here the maximum case has been viewed but in actual practice the change may be only two to one resulting in a six decibel difference in level. The receiver with its linear diode rectifier detector, cannot be expected to improve upon that which is radiated by the transmitter.

Nevertheless, there are circuits which can be incorporated in the audio amplifier system of a receiver which

will compensate for the shortcomings of the transmitter in this respect. Some of them operate on the principle of increasing the audio amplification as greater diode voltages are developed, but at a much higher ratio. Other methods are based on an increasing power output. This latter principle is used in the WLW model and its description follows.

For an explanation of the power operated volume expander reference is once again made to Plate I. In this drawing, 79 is the expander tube. As the symbol indicates, the device is composed of two resistive elements connected in parallel and placed within an evacuated glass bulb. It resembles an ordinary vacuum tube in appearance. For use in the circuit, the two resistive elements in parallel are connected across a portion of the secondary of the output transformer 88. The total of this secondary is designed to work into an impedance of six ohms and the portion across which the volume expander is connected is matched to three ohms. The cold resistance of the expander tube as used here is also three ohms, but at incandescence it rises to approximately thirty ohms. When the resistors are cold, only one-half of the total current flows through six ohm load across the total secondary. An incandescence, however, practically all of the current flows through the six ohm lead. This represents a current change ratio of two to one or a power change of four to one. A four to one power change means six decibels change. This value represents the audio expansion alone, but added to a poss-

ible 20 decibels change in modulation percent gives a total of 26 decibels for the expansion of audio power. While this falls short of the 70 decibels expansion necessary for full and perfect rendition of a musical selection, it does compensate to some extent for the limitations imposed by the transmitter, and the psychological effect is much more pronounced than the additional six decibels should warrant normally.

PART IV

THE L-2 CHASSIS

GENERAL

The L-2 chassis is purely an audio amplifier designed to permit high fidelity reproduction at high and low power output levels. It is well known that the average radio receiver lacks the facilities of high fidelity reproduction at output levels of less than about one watt. When it is recalled that the average radio listener utilizes but about fifty milliwatts of output power for programs in his home, the one watt level seems excessively high. And so it is! Not only does the average listener not care for such outputs normally, but he is obliged not to use them for fear of disturbing his neighbors. Therefore, he is constantly losing or missing a part of every musical program. The power amplifier of the L-2 chassis, with its loudspeaker bank, enables the listener to obtain faithful reproduction at average volume. On the other hand, if the occasion demands, he can use the latent power that exists.

FREQUENCY CHANNEL DIVISION

Due to the physical separation between L-1 and L-2, the output of L-1 is reduced to six ohms impedance and fed into an input transformer (54, Plate II) of L-2 whose primary was also designed for six ohms (with the proper secondary loads). The secondary of the input transformer is divided into two sections: (1) low and middle frequencies, (2) high frequency.

Referring now to Plate II, it can be seen that one secondary of transformer 54 has resistors 29A and 29B in

series connected across it. These two resistors total up to 900 ohms which combined with the 500 ohm load across the other secondary reflects six ohms into the primary, thus matching the output of the L-1 chassis. The low and middle frequency channels are common through the 900 ohm secondary and tubes 40B and 40C which operate in push-pull. In the output circuit of these tubes, across choke 1, the two channels divide.

The drawing shows that the volume control 61A is also connected across choke 1. This is a dual volume control, the arms of which feed the four input grids of the output tubes 41G, 41H, 41I, and 41J which operate in parallel push-pull. They are capable of delivering a maximum output of approximately fifty watts. The condenser 16A and 16B in combination with choke 1, results in a circuit resonate at 25 cycles per second. A response of +20 decibels is obtained at that frequency.

The middle or mezzo frequency channel is taken from across choke 1 through condensers 66A and 66B. These two condensers are in turn connected to the outside terminals of the dual volume control 61B. The arms of this control feed the two input grids of the output tubes 41E and 41F which operate in push-pull with a maximum output of ten watts.

An interesting circuit is the high frequency channel. The high side of the remaining secondary feeds the grid of the first amplifier tube 39B through choke 2. This choke

2, in combination with condensers 8 and 64, make up a filter to cut-off at 12,000 cycles per second. Decision for this cut-off frequency was based on actual listening tests of the receiver. The output of the tube 39B is resistance capacity coupled to the second amplifier tube 40A. In the output circuit of this tube, condenser 10A and choke 3 combine to resonate at 8,000 cycles, per second, with a response of +55 decibels. The choke 3 also permits conversion from a single-ended amplifier to a push-pull input to the output tubes 41C and 41D.

CHANNEL LEVEL CONTROL

The two dual controls 61A and 61B, control the low and middle frequency volumes respectively and volume control 60 adjusts the high frequency channel. By means of these three controls, it is possible to obtain an infinite number of combinations for any desired output response characteristic. It must be remembered that these controls merely adjust the levels of the different channels; the frequency range of each channel remaining fixed by design.

PUBLIC ADDRESS FEATURE

A radio receiver so extraordinary as the one being described in this treatise would not be complete without its public address feature. The pre-amplifier circuit proper consists of five tubes and, referring to Plate II, they are designated as tube 42 microphone input amplifier tube, 39A second amplifier tube, 39C phase inverter, which combined with tube 39A gives push-pull output, and tubes 41A and 41B the push-pull output tubes. This pre-amplifier

channel has a maximum net gain of 85 decibels, which is necessary for use with the crystal microphone 18. The output of the microphone pre-amplifier from tubes 41A and 41B is fed into transformer 58 whose secondary is designed to match six ohms impedance. The microphone relay 19, energized by the total cathode current of tubes 41A and 41B switches the microphone channel output to the input transformer 54. Assuming that it is desired to use the public address system, the operation is as follows:

A small button is pressed in the handle of the microphone which shorts the relay coil (normally energized). Now with switch 38 in the position as shown, the relay being de-energized makes contact as shown, and the radio program is cut-off allowing the microphone channel output to be fed to all three regular audio channels. If the switch 38 is thrown in the other direction, speech, singing, or other sound affects may be blended with the radio program. Its flexibility permits almost any use that could be desired.

There is yet another interesting feature of the public address channel. Due to the exceedingly high gain resulting from tubes 42, 39A, and 39C, any vibration of them causes a microphonism of the tubes themselves. This phenomenon was so pronounced that some method of elimination had to be employed. The problem was solved by mounting all three of those tubes on a heavy steel block which in turn was mounted to the chassis by means of flexible

rubber bushings. The high inertia furnished by the steel block combined with the rubber bushings allows no transmission of vibration to the tubes.

PART V

POWER SUPPLY

THE L-3 CHASSIS

Because of the power demanded by the L-2 chassis, a separate power supply was deemed advisable, and the L-3 chassis was designed to furnish the necessary d-c power. This chassis is split up onto two separate power supply circuits, one for the low, middle, and microphone channels, and the other for the high frequency. These were designated LPS and HPS respectively.

The LPS circuit is composed of a power transformer, two rectifier tubes, filter choke, and filter condenser. The power transformer of this circuit was designed for a rating of 400 watts at 375 volts output. A voltage regulation of five per cent also was demanded. In Fig. 17 it is shown as transformer 8. The filter choke 4 was designed to have an inductance of 8 henries with 300 milliamperes current flowing through its winding. The filter input condenser 1A has a rating of 35 microfarads at 400 volts. The four filter output condensers 2A, 2B, 2C, and 2D each have a rating of 40 microfarads for the output filter condenser. Such high capacity was necessary to absorb surges which were often induced due to sudden output changes.

The HPS circuit is somewhat more simple as less demand is present. The power transformer 9 was designed for 150 watts at 375 volts output. For a filter choke in this circuit, one of the middle frequency range speaker fields is used. Only 80 microfarads of output filter is used

since the surges are not so great as they were in the LPS circuit.

Output of the two circuits was fed be cable to female plugs which in turn connected to the L-2 chassis.

THE L-4 CHASSIS

As was stated above the d-c field supply for one of the middle frequency range speakers is furnished by the HPS circuit of the L-3 chassis. The compliment of this speaker has its field supplied from the power unit of the L-1 chassis.

Field supply for the one low frequency speaker and three high frequency speakers had to be provided by a separate chassis. The L-4 chassis whose circuit diagram is shown in Fig. 18 was designed for this purpose. Although simple, it had to furnish 65 watts of d-c at 240 volts. Since the power demand on its circuits was steady, very little precautions had to be taken for regulation. Nevertheless, the design of the transformer is conservative.

An auxilary winding furnishing two amperes at six volts had to be provided for the four dial lights which are used to illuminate the side control panels which will be discussed under the part on summation.

The output of the L-4 chassis was fed through cables and plugs to the different speaker fields.

All power transformers were designed to operate from 110 to 120 volts 60 cycle power supply. It is interesting

to note that the complete receiver consumed 540 watts of power when no signal was received, and 600 watts at maximum audio output. Overall efficiency is, therefore, 75/615 x 100 or 12.2 percent.

PART VI

SPEAKER AND CABINET CONSIDERATIONS

LOUDSPEAKER COMPLIMENT

The bank of loudspeakers used for this receiver consists of one 18 inch auditorium type which reproduces the bass frequencies, two 12 inch speakers which reproduce the middle frequencies, and three special diaphram type, commonly known as tweeters, which reproduce the high frequencies. These speakers, are of course, the most expensive that can be purchased. Being of good design, their reproduction is of the best quality obtainable. As a matter of interest, the 18 inch speaker alone weighs 85 pounds. None of the speakers are fixed in the cabinet for shipment. Their combined weight prohibits such procedure. Where more than one speaker is used for one channel, proper phasing of the voice coil circuits had to be observed. It is obvious that if one cone is moving in one direction and another in the opposite direction, at the same instant, good reproduction will not be obtained. The two cones, must, therefore, be moving in the same direction simultaneously.

CABINET SELECTION

The selection of the cabinet for the WLW Model receiver was based on two primary decisions; (1) it had to be large enough to house all the necessary equipment, and (2) although large, it had to be attractive. As a consequence, the style is modernistic in the extreme. There are seven different kinds of wood used in its construction, the most beautiful being the band of Macasar

Ebony at the bottom. Some idea of its immense size can be obtained from the photographs. The young lady standing beside the cabinet is of medium size. The dial furnishes another clue; being 12 inches in diameter.

PART VII

SUMMATION

RESUME OF FEATURES

In one great assembly, with a weight totaling better than 475 pounds, this is a receiver capable of continuous tuning from 540 kilocycles to 18,000 kilocycles, capable of reproducing the complete scale of audio frequencies, and capable of public address volume to handle crowds of 10,000 people.

The two accompanying photographs showing front and rear views are designated Plates III and IV respectively. Plate IV is especially interesting in that it shows arrangement of all the equipment. In the center on the upper shelf can be seen the L-1 chassis. The L-2 chassis is located at the lower left corner, and the L-3 chassis at the lower right. The L-4 chassis can be seen on the middle shelf. The large speaker is located in the center at the bottom with the two mezzo speakers on each side. The large speaker is suspended from the middle shelf on rubber supports, and due to its great driving force at low frequencies, is spaced back from its baffle by one-fourth inch. This precaution was necessary to prevent cabinet resonance at a particular frequency, and to prevent vibrational transmission at other frequencies, to the chassis, resulting in microphonism which produced a low frequency rumble. The three "tweeters" are also mounted on the middle shelf and are focused in three different directions. The two aluminum housings at the upper right and left contains the side panel controls. From the rear of each can be seen flexible dental cable connecting them with the L-2 chassis. These

are the channel and microphone controls. Plate III shows these two controls. Attention is directed to control panel at left. The bottom knob controls the bass channel, the center controls the mezzo channel, and the upper controls the treble channel. In the right control panel, the upper knob controls switch 38 of Plate II and the lower one controls the microphone channel output.

The four chassis and all their component parts that could be are chromium plated. Transformer cores, tubes, and speaker frames are finished in black.

The writer enjoyed every minute spent on the creation of this receiver and welcomed the responsibility of making it a commercial possibility.

ACKNOWLEDGEMENT

The writer wishes to express his appreciation to his colleagues without whose help this development could not have been accomplished so rapidly. He is especially indebted to Mr. E. J. H. Bussard, Chief Development Engineer, The Crosley Radio Corporation for his moral support during the long hours of development work.

Further appreciation is expressed to the Crosley Radio Corporation who furnished the writer with the necessary data with which to write this thesis.

The writer is also deeply grateful to Mr. Garland P. Baker, who typed this entire thesis.

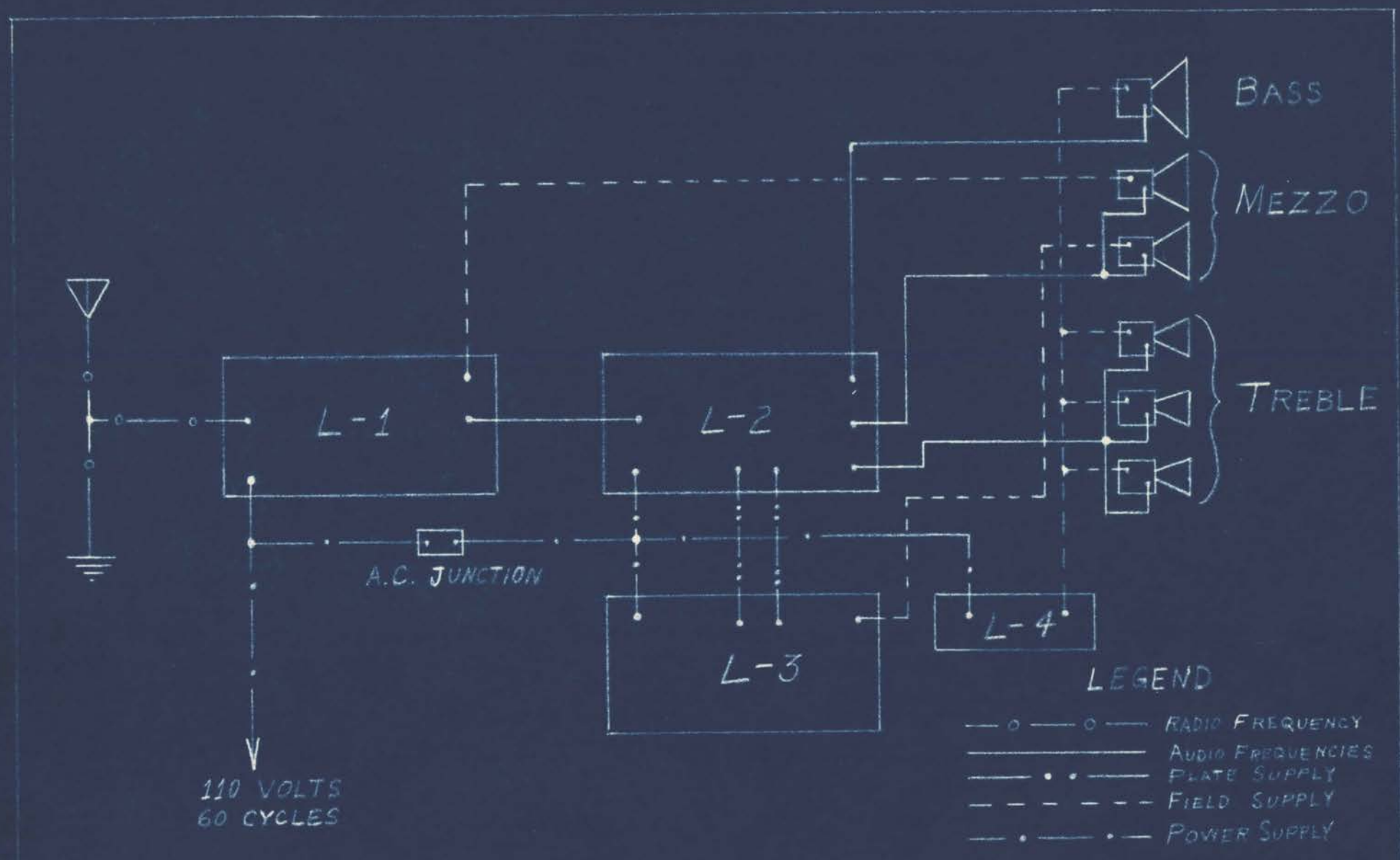


FIG. 1

12-26-31
APR

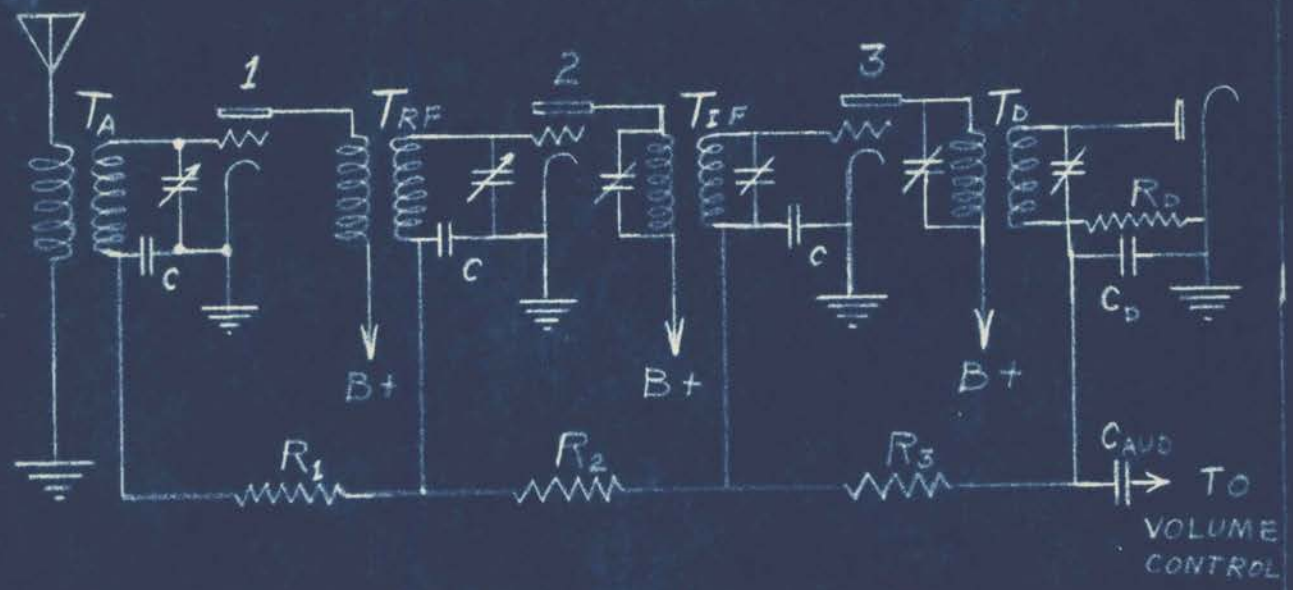


FIG. 2

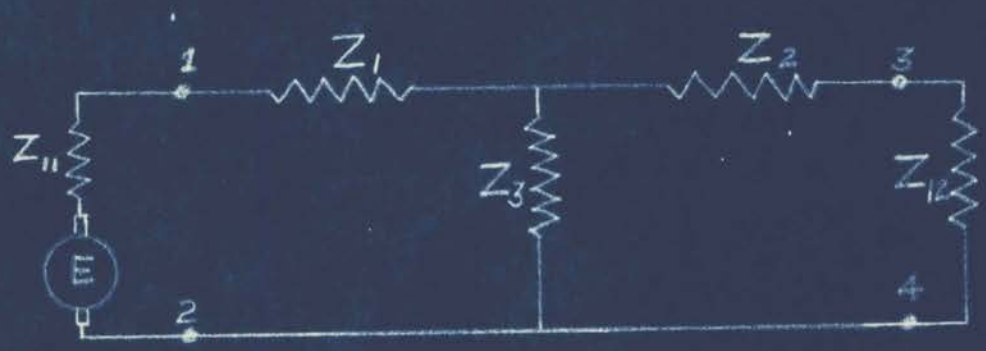


FIG. 3

1-15-38

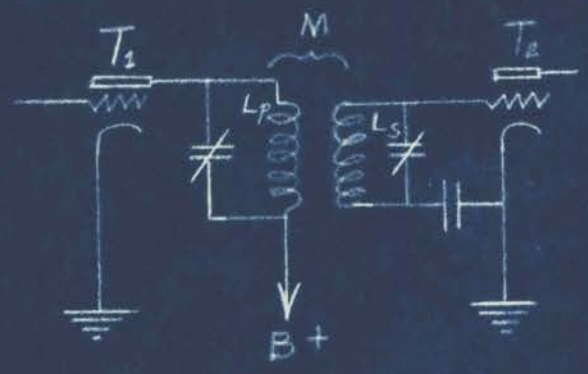


FIG. 4

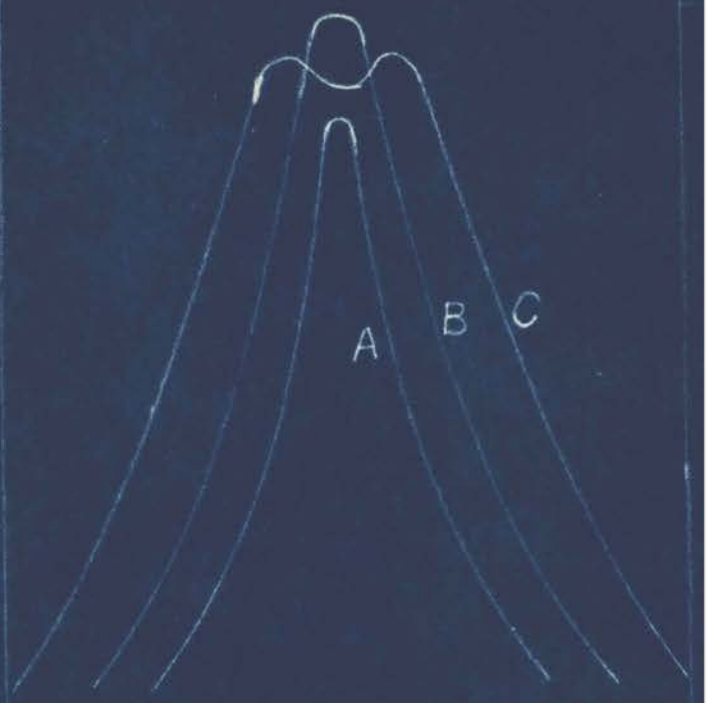


FIG. 5

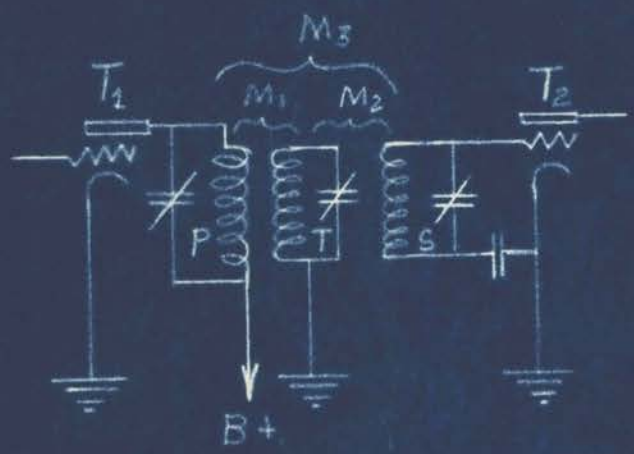


FIG. 6

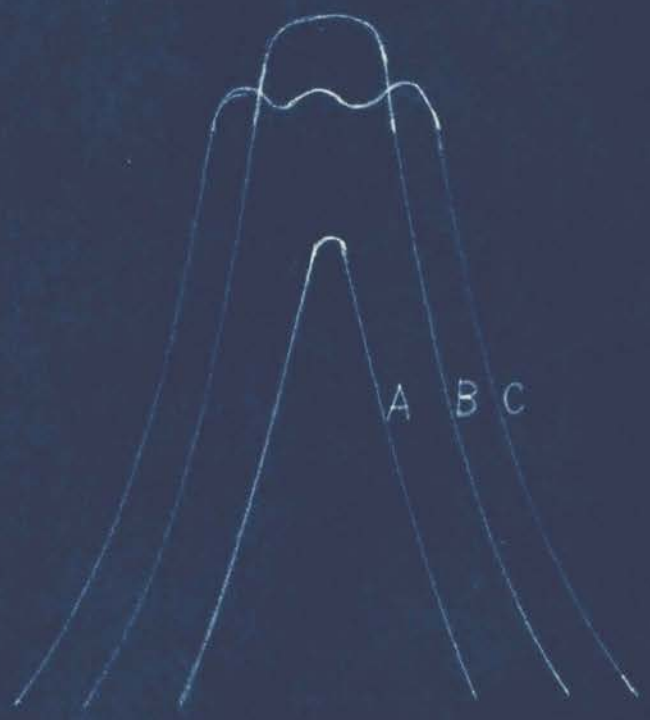
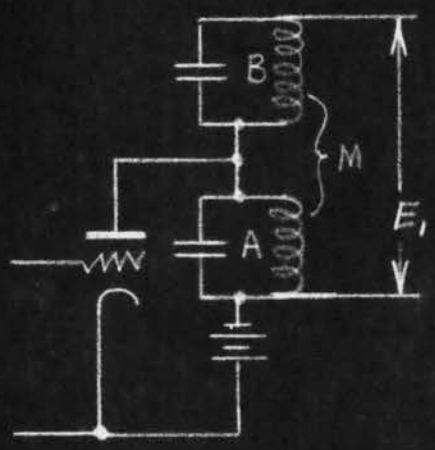


FIG. 7

1-23-38



CIRCUITS A AND B BOTH TUNED TO f

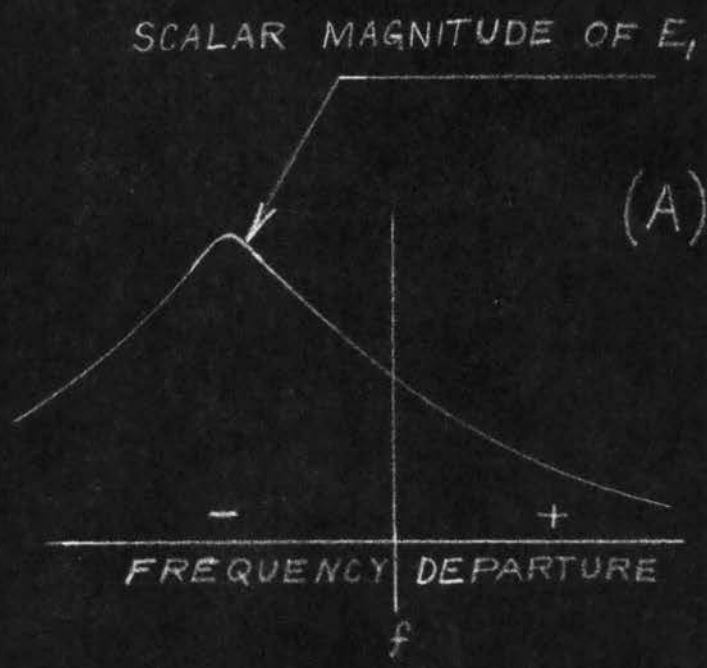
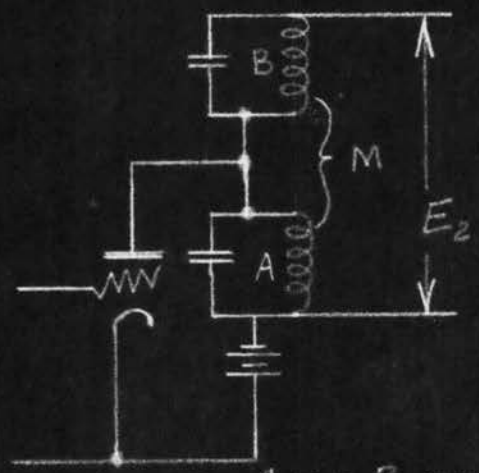


FIG. 8



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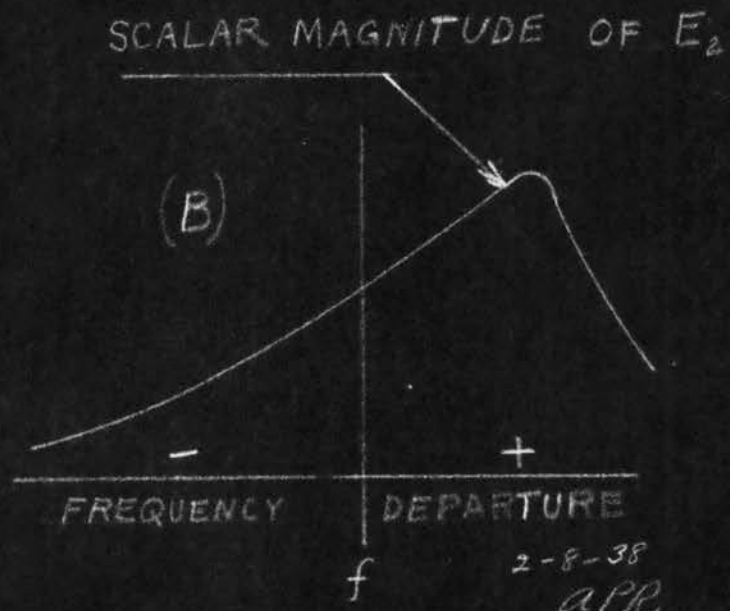


2-8-38
aPR

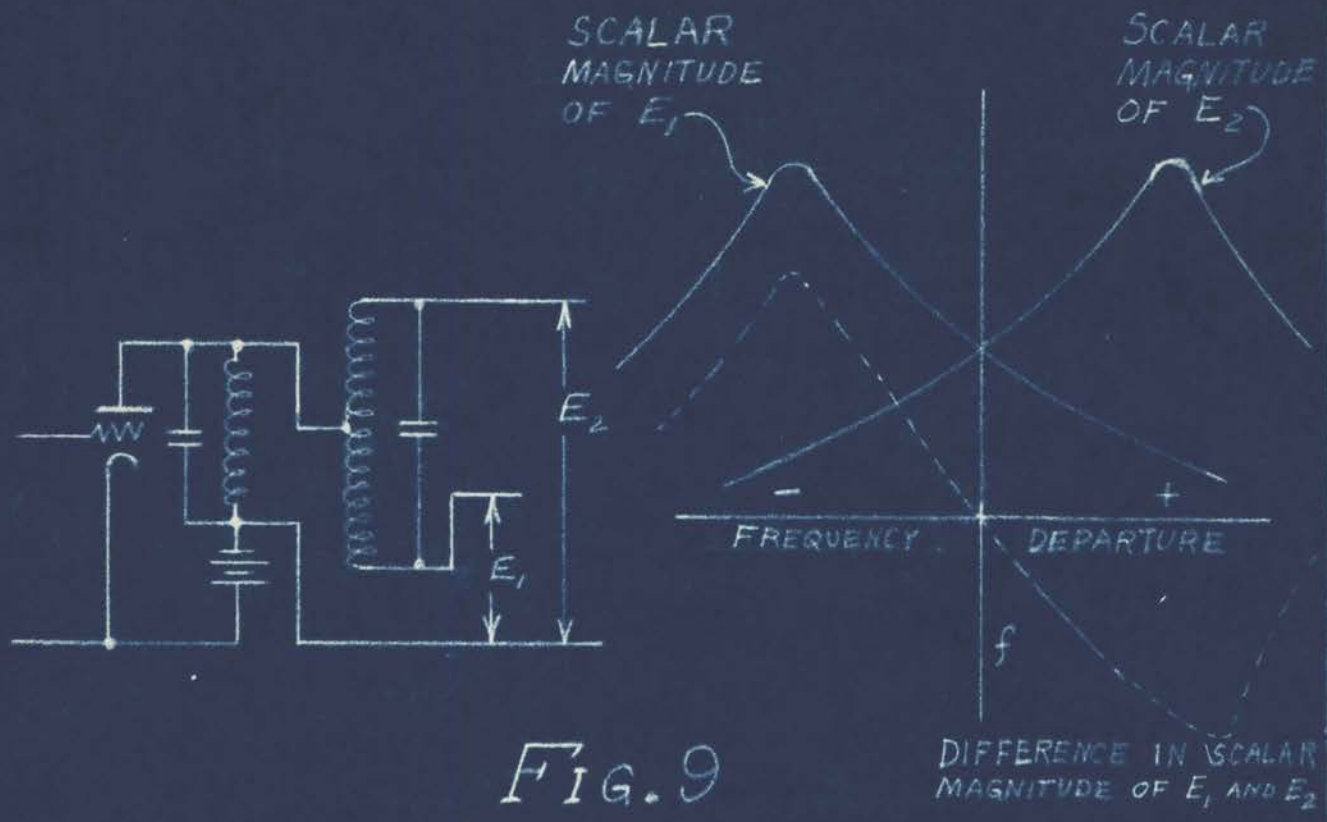
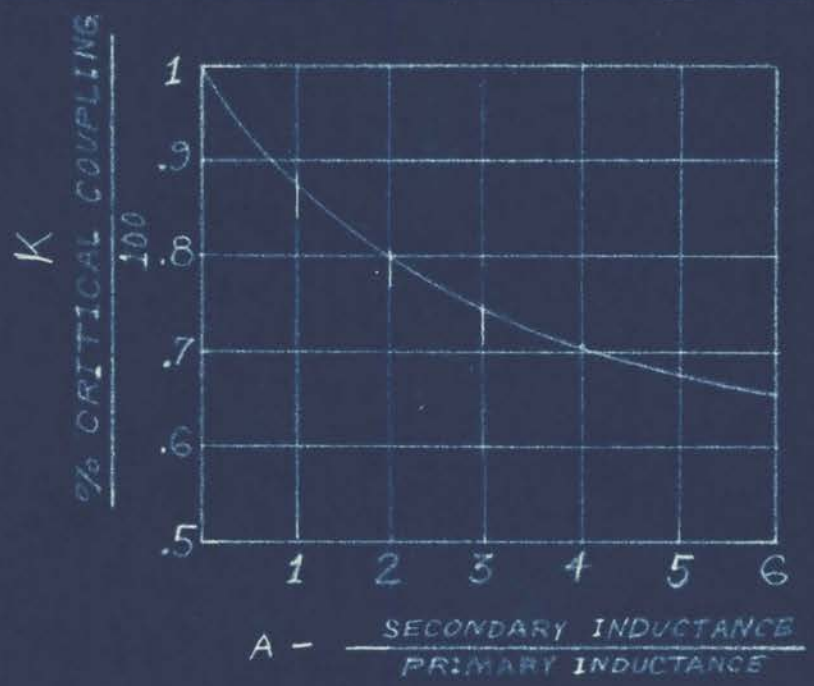


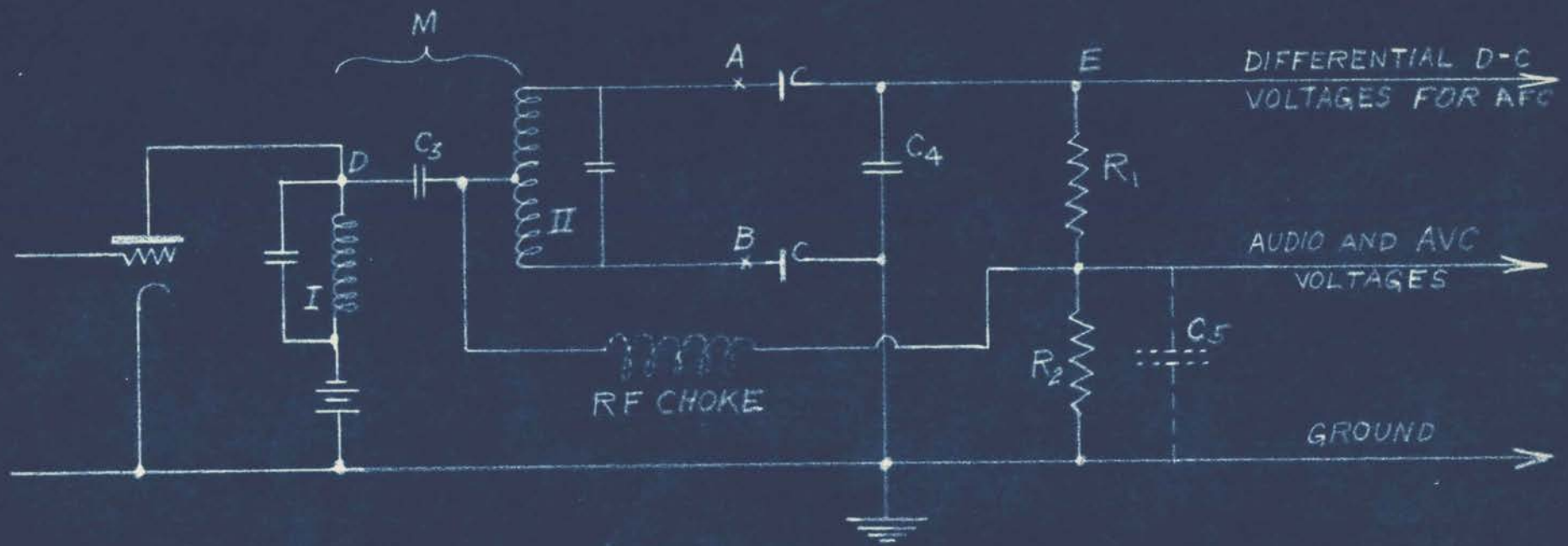
FIG. 9



OPTIMUM COUPLING IF LINEAR DETECTORS ARE USED

FIG. 10

2-23-38
AHL



IF THE RF CHOKE AND THE CONDENSER C_5 ARE NOT USED AN RF CHOKE MUST BE INCLUDED IN ANY EXTERNAL CONNECTION TO THE "AUDIO AND AVC VOLTAGE" POINT.

FIG. 11

4-5-38
APR

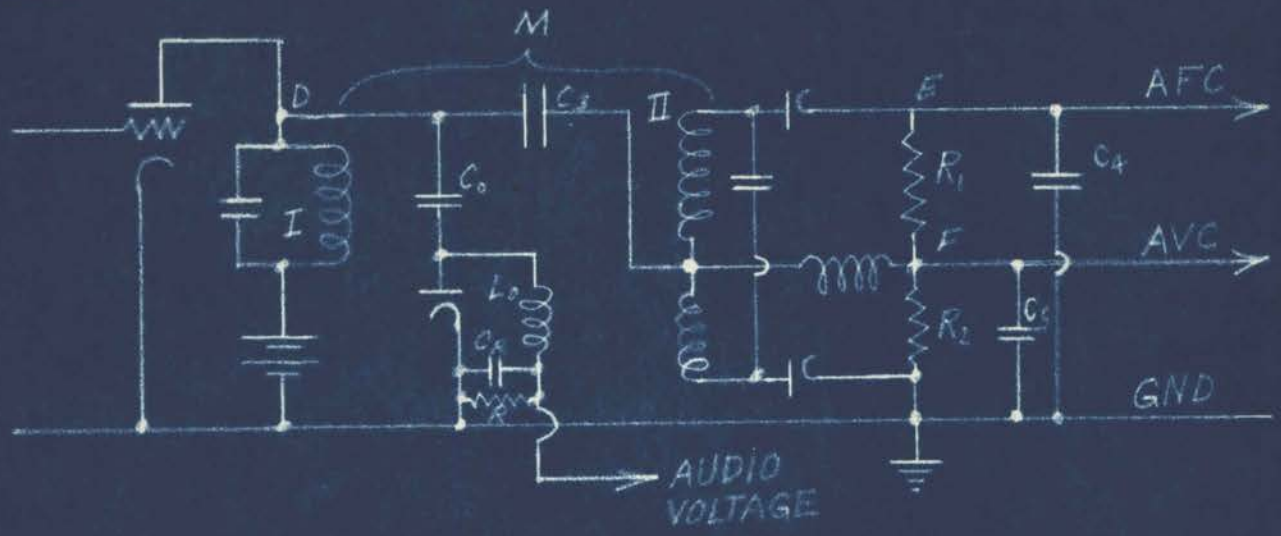


FIG. 12A

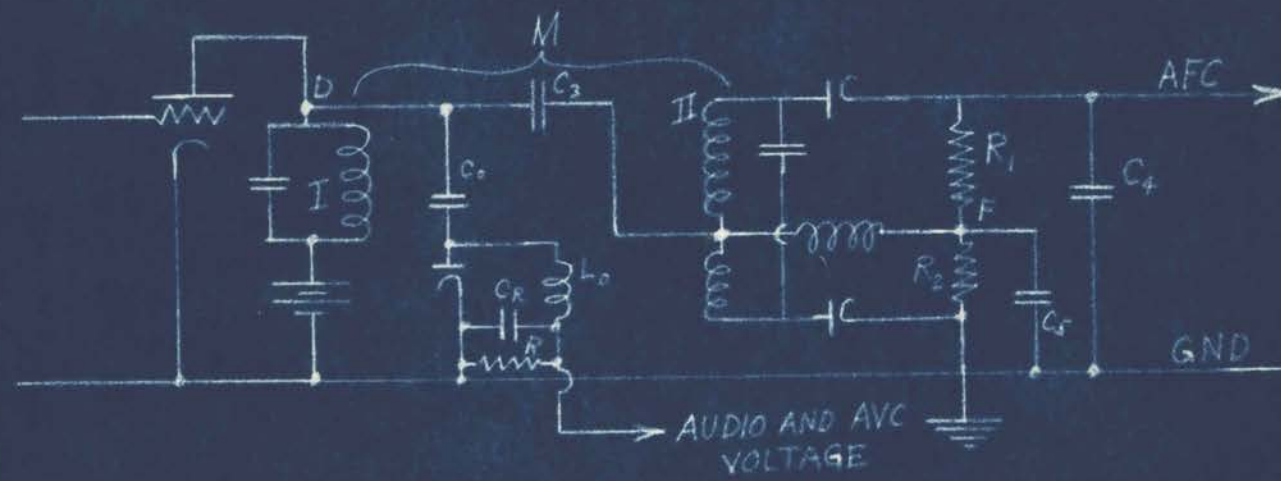


FIG. 12B

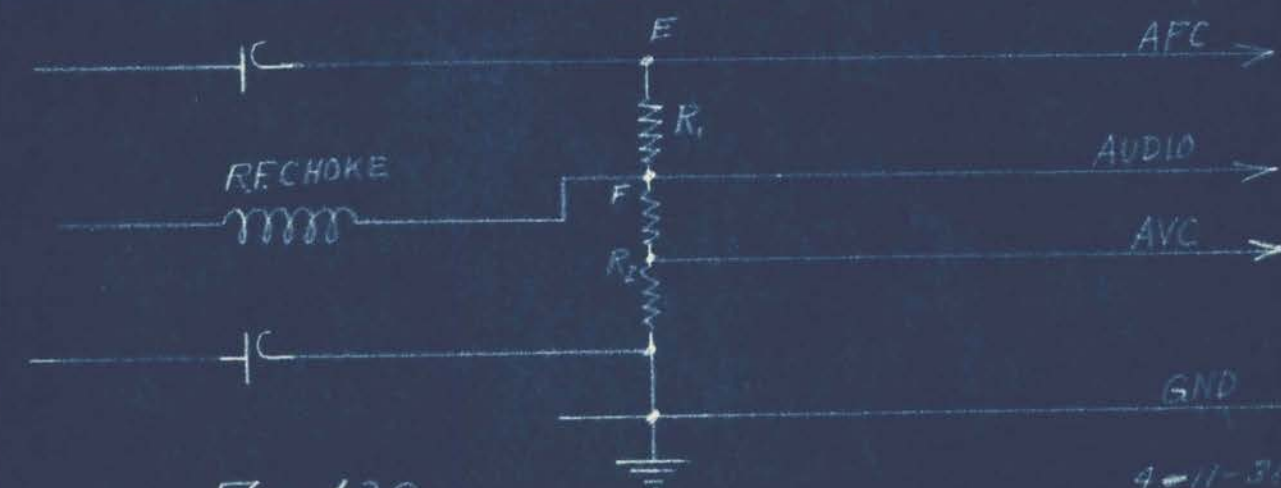


FIG. 12C

4-11-34

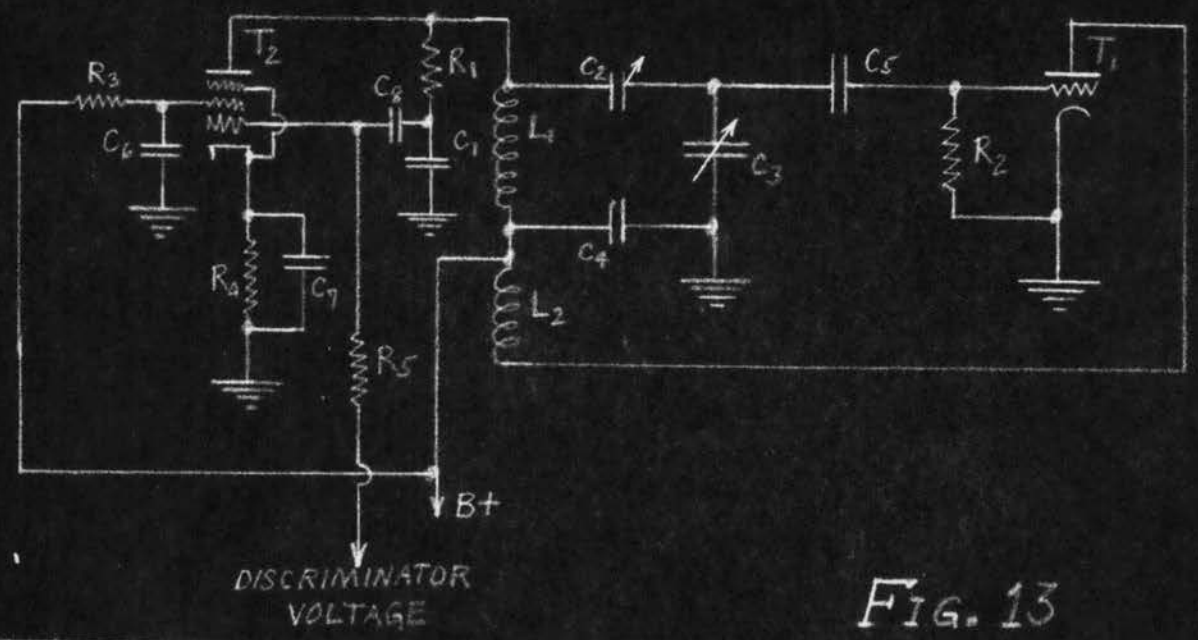


FIG. 13

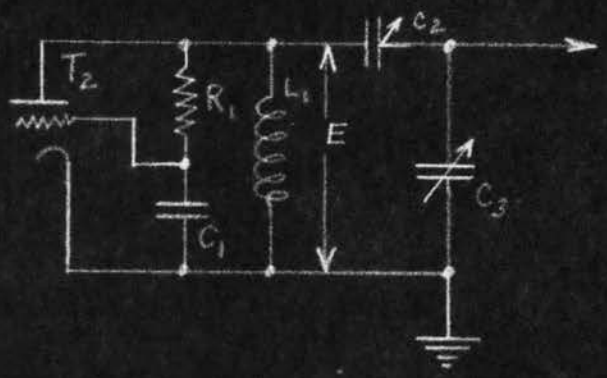


FIG. 14

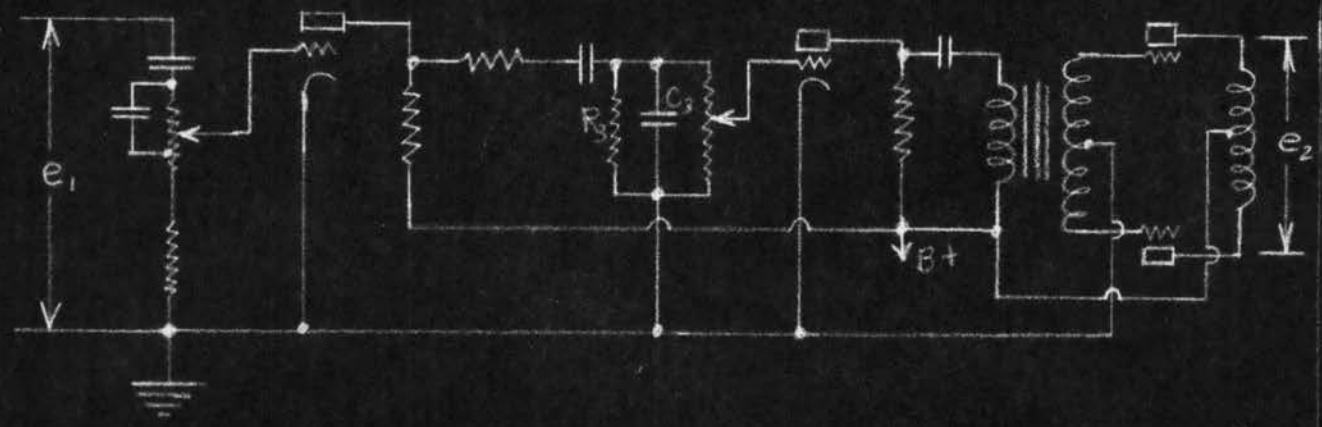


FIG. 15B

4-11-38
APR

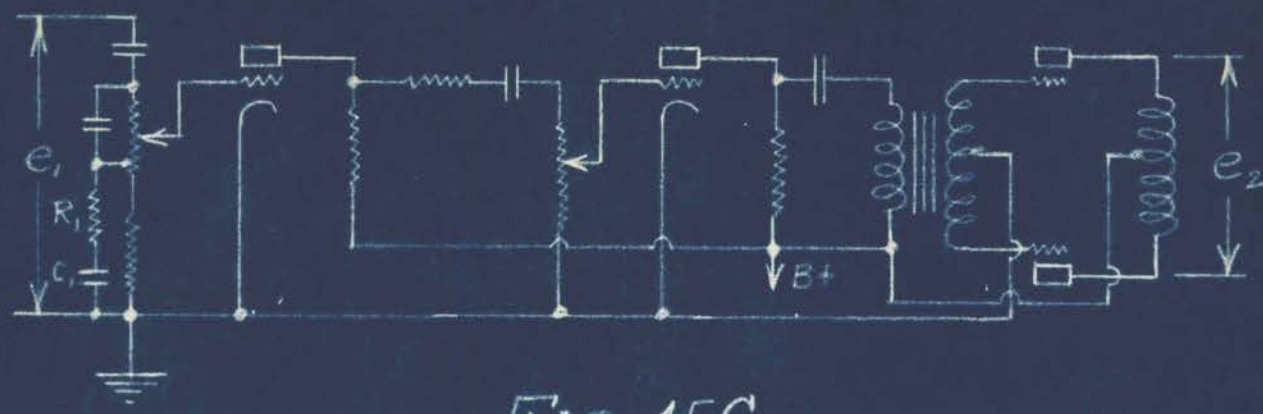


FIG. 15C

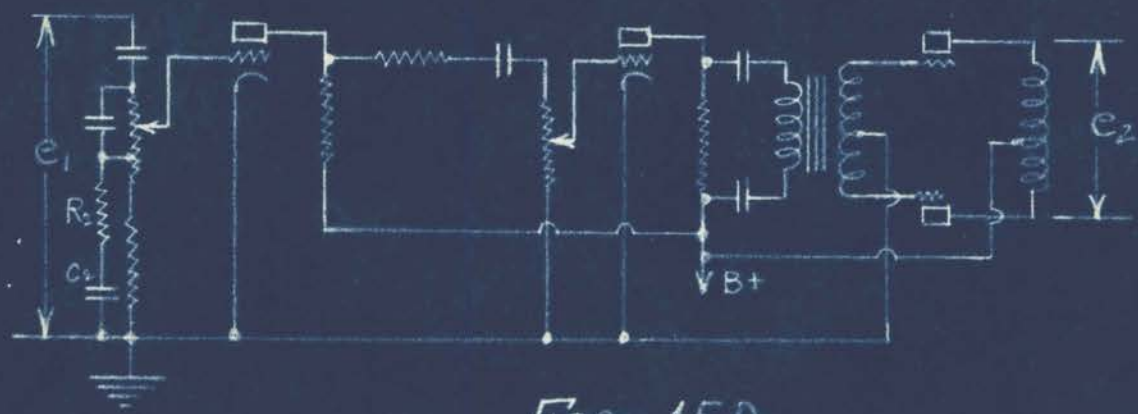


FIG. 15D

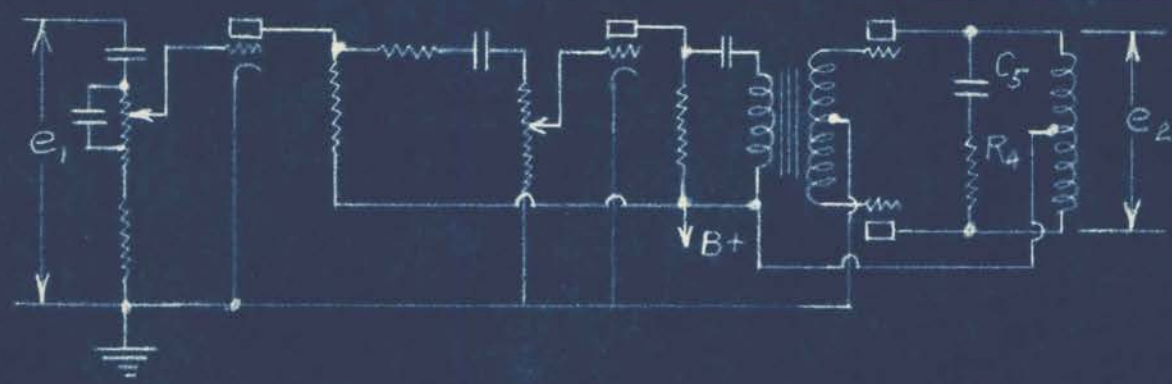


FIG. 15E

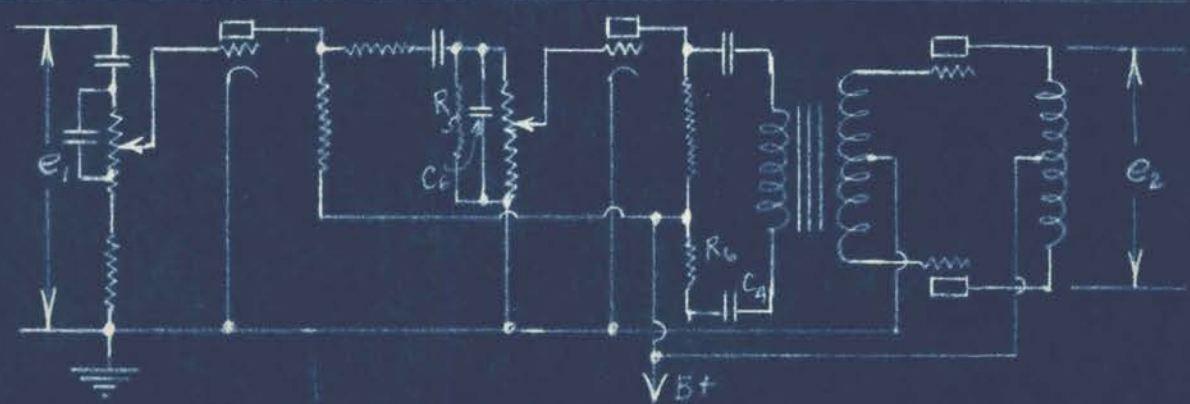


FIG. 15F

4-11-38
DPR

CROSLEY RADIO CORPORATION

ENGINEERING DEPARTMENT

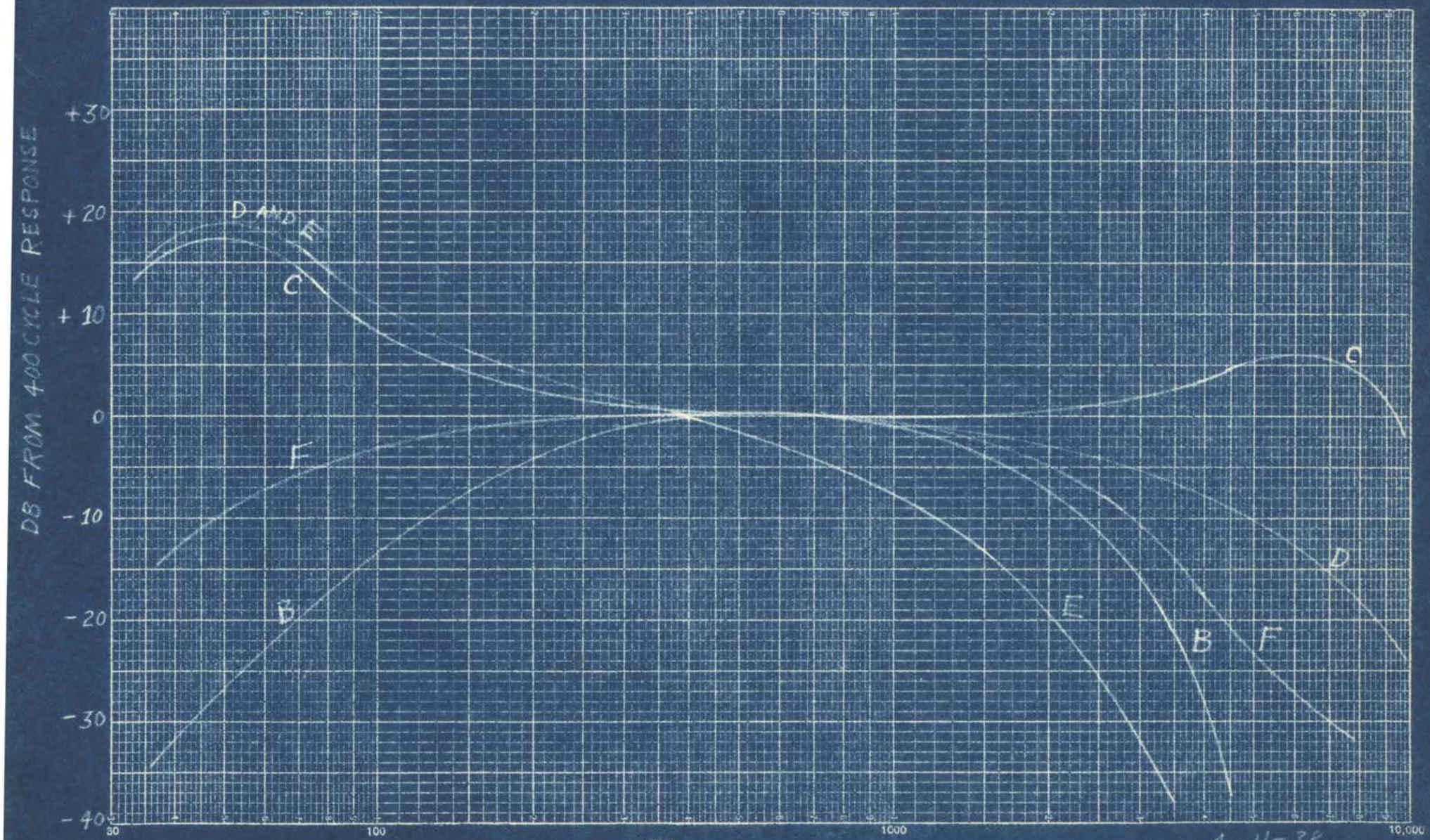


FIG. 16

4-11-38
APK

FREQUENCY CYCLES PER SECOND

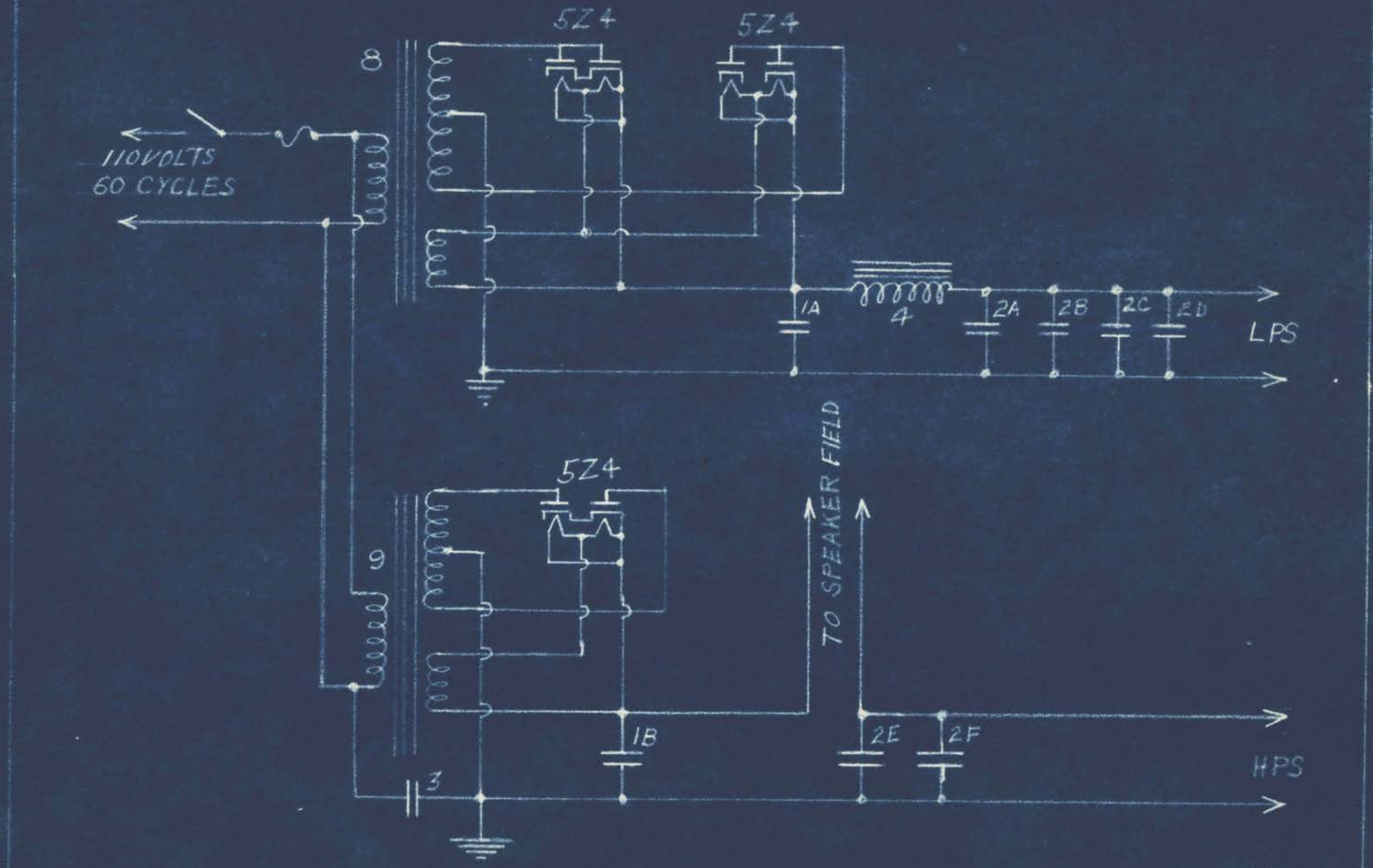


FIG. 17

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aPR

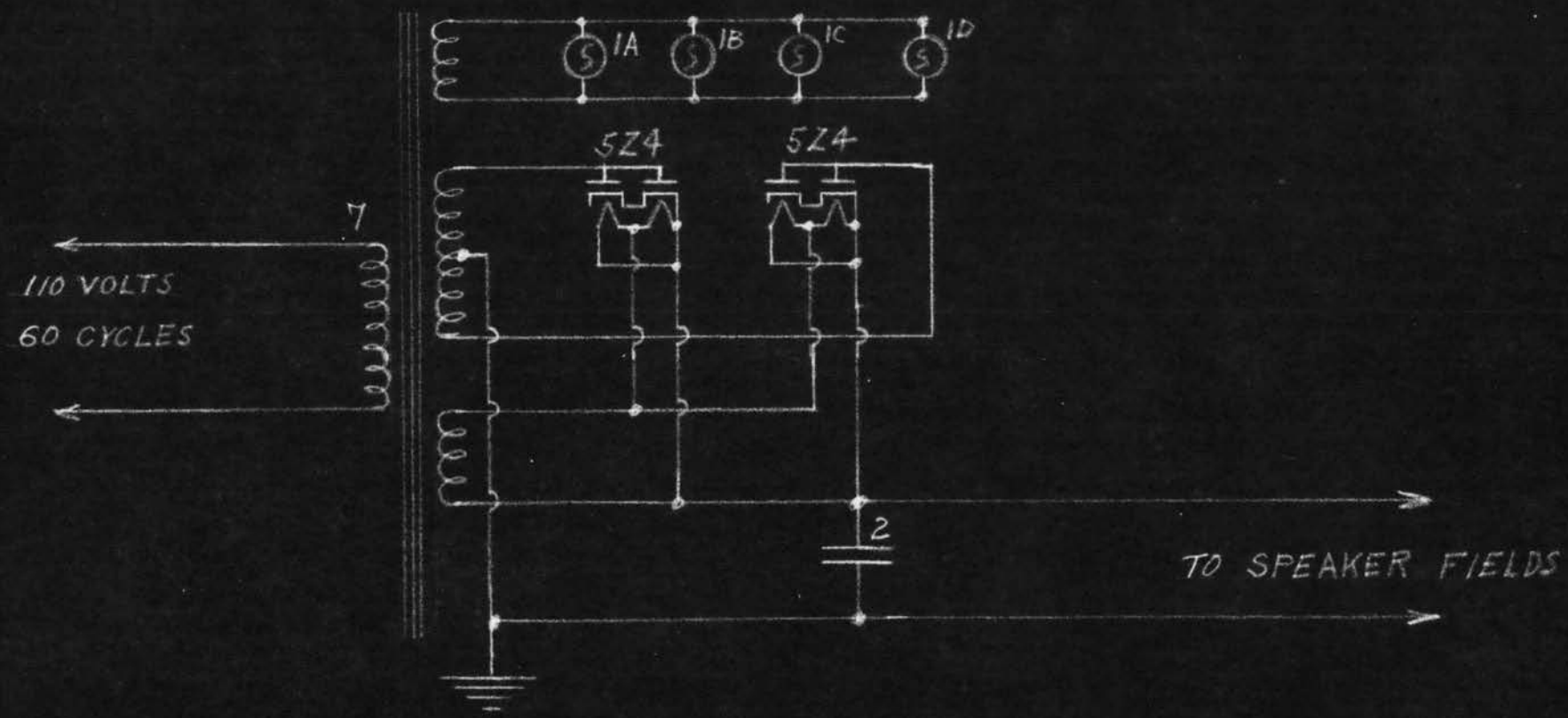


FIG. 18

4-11-38
ORL

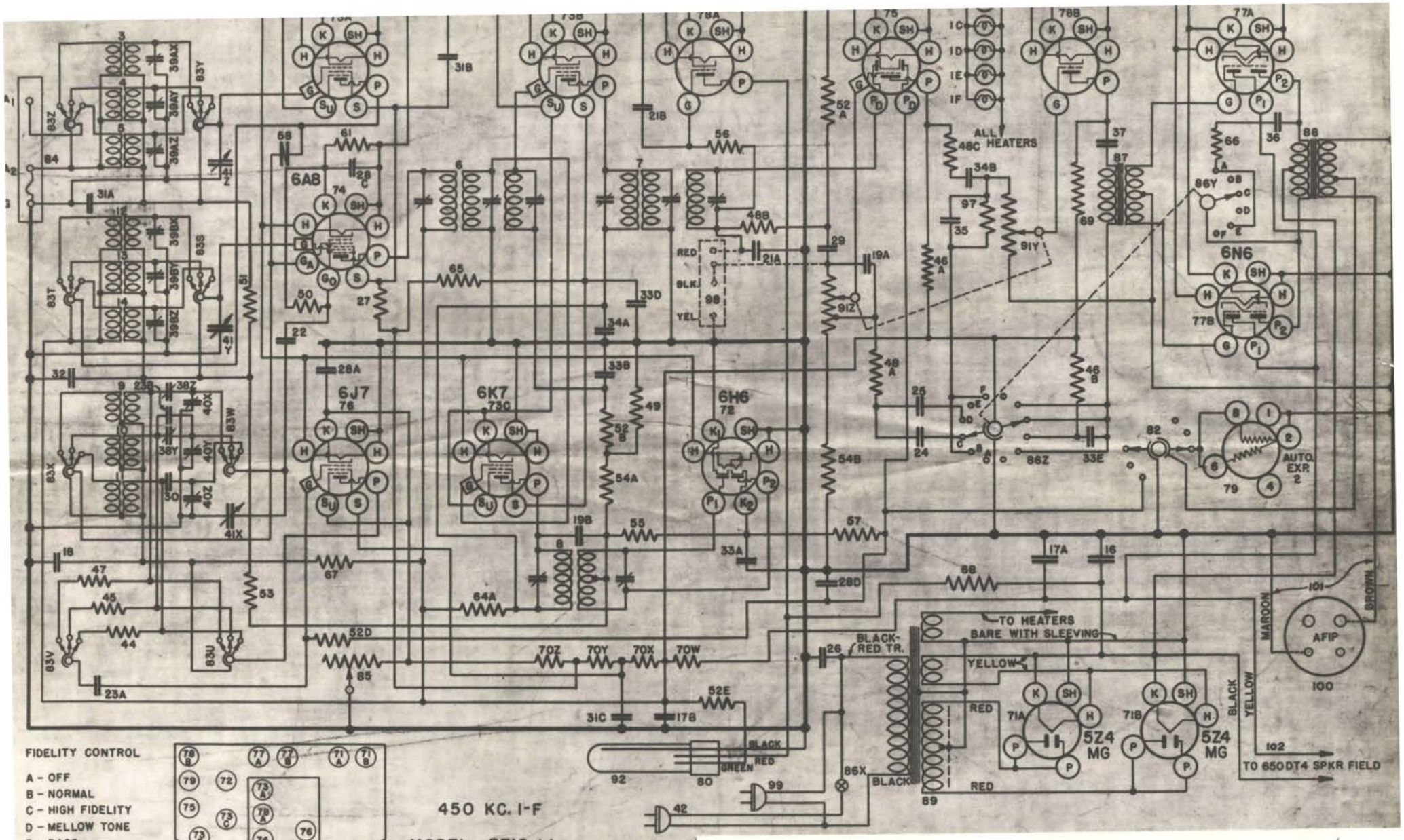
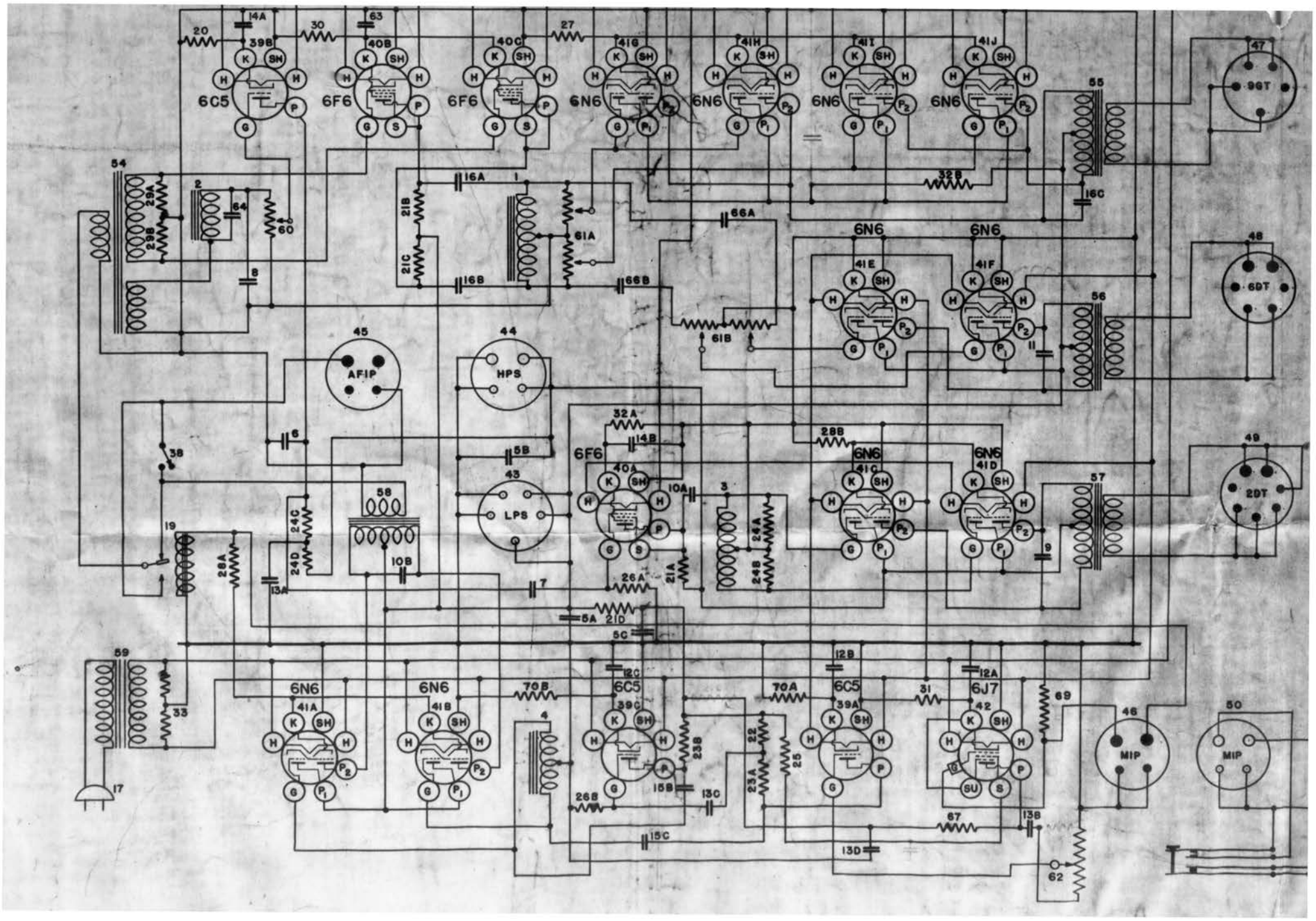

6A8
6J7
6K7
6H6
6N6
5Z4 MG
5Z4 MG
AUTO. EXR. 2
AFIP
ALL HEATERS
TO HEATERS BARE WITH SLEEVING
BLACK-RED TR.
TO 650DT4 SPKR FIELD
FIDELITY CONTROL
A - OFF
B - NORMAL
C - HIGH FIDELITY
D - MELLOW TONE
E - BASS
F - NOISE REDUCING
450 KC. I-F
MODEL - 3716 LI

PLATE I


6C5
6F6
6F6
6N6
6N6
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6N6
6N6
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6F6
6N6
6N6
6N6
6N6
6C5
6C5
6J7
AFIP
HPS
LPS
MIP
MIP
9GT
6DT
2DT

TYPED BY

GARLAND P. BAKER